

Master Thesis

Design and Optimization of a Broadband Electromagnetic Kicker for the Stochastic Cooling System of the Antiproton Decelerator at CERN

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Abstract

The response of the pickups and kickers of the stochastic cooling system of CERN's Antiproton Decelerator is investigated by means of analytical computations and numerical simulations.

A novel class of stochastic cooling pickups and kickers, based on striplines coupled with pulse inverters is presented. The frequency domain behavior of this class of devices is analyzed analytically and by means of numerical simulations. Designs of kicker arrays of devices of this class are compared to the currently used designs and to a slotline array kicker design.

In Gedenken an Fritz Caspers.
(†12. März 2025)

Ich habe unendlich viel von Dir gelernt. Danke dafür.

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1 Introduction

CERN, the European Organization for Nuclear Research, operates the world's largest complex of particle accelerators. An illustration of the complex is shown in Fig. 1.1. This thesis is related to the Antiproton Decelerator (AD), which is one of the few machines at CERN that decelerates charged particle beams, instead of accelerating them. The AD is located inside CERN's Antimatter Factory, where several experiments¹ investigate the properties of antimatter.

The experiments need to be supplied by low energy antiprotons. Antiprotons are produced by delivering a high energy proton beam from the Proton Synchrotron (PS) onto an iridium target. A wide variety of different types of particles are produced in the target, one of them being antiprotons. The antiprotons which are produced in the target are focused and guided into the AD, a decelerator ring of 182.43 m circumference [6]. In the AD, the antiprotons are decelerated from a momentum of 3.57 GeV/c to 100 MeV/c (Fig. 1.2). This corresponds to a deceleration from the velocity of approximately 97% of the speed of light, down to a velocity of approximately 10.6% of the speed of light. The 100 MeV/c antiprotons are then transferred to the Extremely Low Energy Antiproton ring (ELENA), where they are further decelerated.

The stochastic cooling system of the AD plays a crucial role in reducing the beam emittance, both longitudinally and transversely. This is necessary to reduce the size of the injected antiproton beam and to give enough space to the beam during the deceleration process. Emittance increase leads to losses and therefore fewer antiprotons available for experiments. The stochastic cooling system of the AD is active on two plateaus of the momentum program, which is shown in Fig. 1.2. On the first two plateaus, the deceleration is paused and the stochastic cooling reduces the momentum spread and the transverse emittance. On the third and fourth plateau, the emittance is reduced by electron cooling, which is not further discussed here.

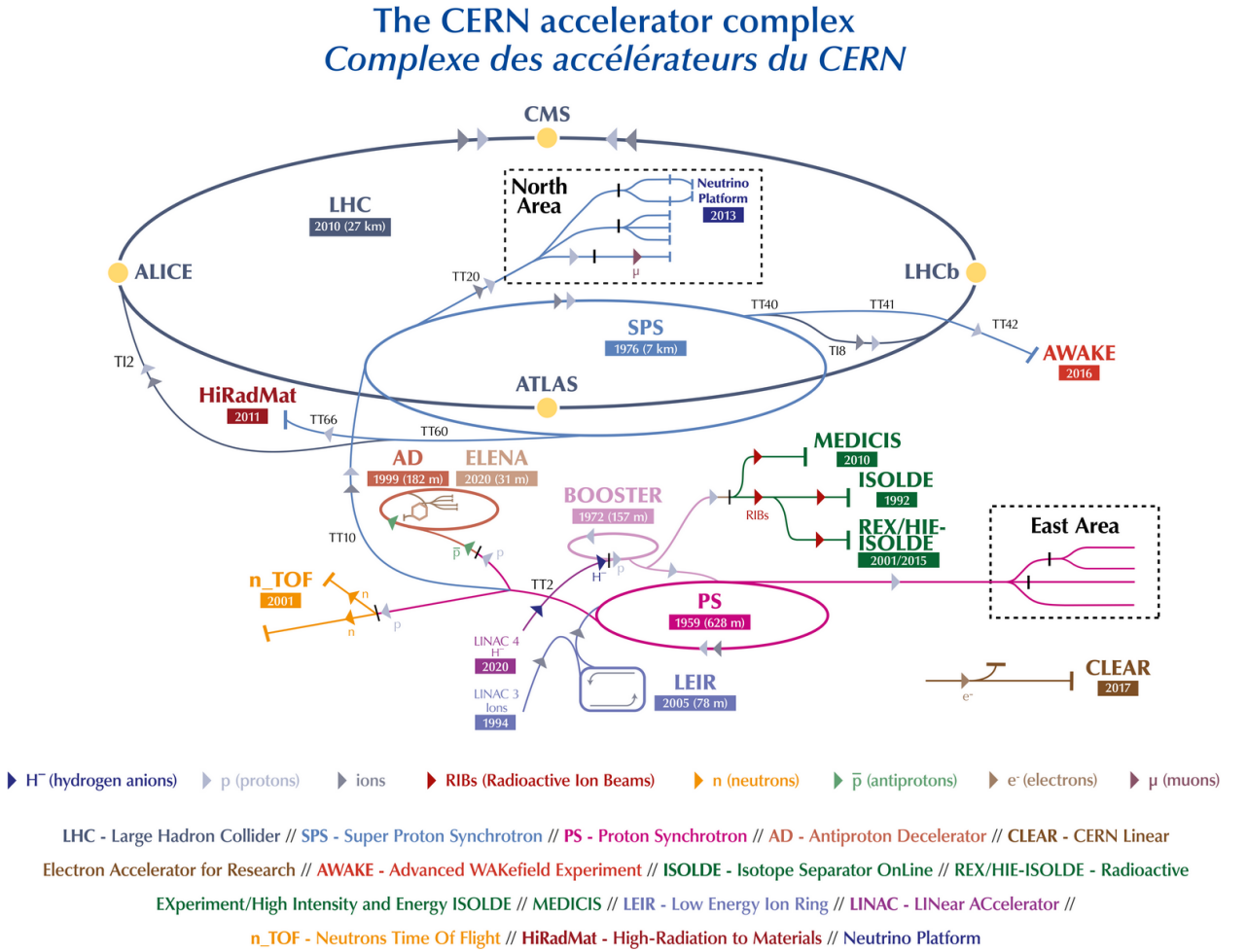


Fig. 1.1: The CERN accelerator complex (cropped from [7]).

¹The experiments performed in CERN's Antimatter Factory: AEGIS [1], ALPHA [2], ASACUSA [3], BASE [4], GBAR [5]

A stochastic cooling system can be viewed as a feedback system, that aims to reduce the emittance of a charged particle beam. Either the longitudinal emittance, the transverse emittance² or both are reduced. An illustration of the stochastic cooling system of the AD with its two pickups and two kickers is shown in Fig. 1.3. The transverse displacement of the particles relative to the main beam axis and their velocity are measured by electromagnetic devices called pickups. The measured signals, usually in the microwave frequency range, are processed in the analog domain, amplified and then fed to an electromagnetic device called kicker. The kicker alters the momentum of the charged particle beam such that, over a certain amount of turns, the longitudinal and transverse emittances are reduced.

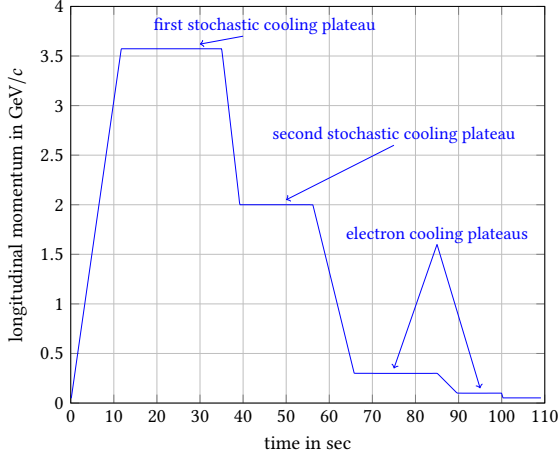


Fig. 1.2: Momentum program used in 2024 for the AD at CERN.

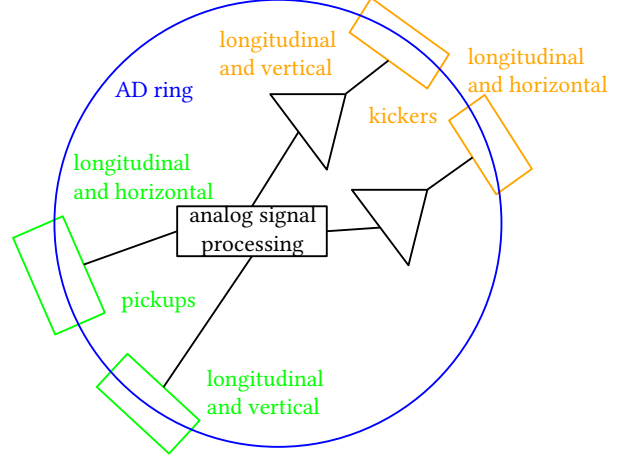


Fig. 1.3: Illustration of the stochastic cooling system of the AD.

The theory behind stochastic cooling is by no means straightforward. In fact, van der Meer shared the 1984 Nobel Prize [8] for his work on stochastic cooling, as this technique was an important enabler to having high energy proton-antiproton collisions in CERN's Super Proton Synchrotron (SPS). A series of experiments with high energy proton-antiproton collisions in the SPS lead to the discovery of the W and Z boson. The W and Z boson are the particles which mediate the weak interaction.

The two pickups and two kickers of the stochastic cooling system of the AD are in operation since the 1980's. The stochastic cooling system is critical for the performance of the AD and as the vacuum tank of the horizontal kicker had leaks in its water cooling circuit in 2001 and 2021 [6], an effort was launched to study the currently used designs and to compare them to alternative designs which could be used for the construction of spares of improved performance. This work is part of this effort.

This work deals with electromagnetic kickers used for stochastic cooling. These are tightly related to pickups due to the reciprocity of Maxwell's equations and thus pickups are also discussed. The details of stochastic cooling are not required to evaluate the performance of these devices. Reference is therefore made to the literature [8], [9] covering theory of stochastic cooling. For the scope of this work, it is sufficient to regard the stochastic cooling system as a feedback system, that operates in a certain microwave frequency band. The pickups are the sensors and the kickers are the actuators of the control system. Both have a certain frequency response which of course should cover the desired frequency band. The operational frequency band of the stochastic cooling system of the AD spans from approximately 900 MHz to approximately 1.8 GHz.

Chapter 2 lays the theoretical foundations of this work. Important theorems and formulas for the electromagnetic devices subject of this work are introduced in Section 2.1. Sections 2.2 to 2.5 provide explanations of the electromagnetic structures subject of this work. Qualitative explanations and derivations of analytical approximations of the working principles are provided whenever it serves the understanding of the performance limitations. References to relevant literature are given whenever appropriate and beneficial. Moreover, some considerations on the numerical computation of the electromagnetic devices subject of this work are discussed.

²The stochastic cooling system of the AD reduces longitudinal, horizontal and vertical emittance.

A tool for computing the so-called longitudinal and transverse beam voltage from time-domain solver results in the CST Studio Suite®[10] is presented. The tool is used for numerical simulations throughout Chapters 3 and 4.

In Chapter 3, the AD stochastic cooling kickers and pickups are investigated by means of numerical simulations using the CST Studio Suite®. Each AD stochastic cooling kicker and pickup is an array of 24 individual elements. The pickup and kicker arrays, respectively, have specific differences that affect their performance on the second momentum plateau. This is taken into account in the investigation. Finally, the simulation results are compared to measurements from 1987 by Autin *et al.*[11].

The main part of this work is Chapter 4, where alternative designs which could be used in an improved design of an AD stochastic cooling kicker are investigated and optimized. One novel class of stripline-type kickers is introduced in Section 4.1, which uses broadband pulse inverters instead of $\lambda/2$ long delay-lines in the design. This modification allows to build stripline arrays of the multi-loop type of improved bandwidth and peak value performance. Moreover, a slotline design is investigated which is particularly simple to fabricate while delivering improved performance, compared to the currently installed kickers.

The conclusion is presented in Chapter 5, and the appendices provide detailed derivations of the formulas used throughout this work.

2 Charged Particle Beam Pickups and Kickers

This chapter deals with the theory, necessary for the investigations and simulations in the other chapters. General electromagnetic theory of pickups and kickers is provided in Section 2.1. Sections 2.2 and 2.3 show the properties of the two topologies of kickers and pickups, that are used in this work.

The behavior of arrays of multiple individual pickup or kicker elements is analyzed in Section 2.4, ending in the important fact that the response of an array of pickups or kickers is determined by multiplying a frequency and β dependent array factor to the response of a single pickup or kicker element.

Performance considerations regarding the phase response and bandwidth of kickers and pickups are discussed in Section 2.5 and in the last section Section 2.6, aspects of numerical computation of kickers and pickups is are discussed. A post-processing template for CST Studio Suite® is presented which allows the computation of longitudinal and transverse beam voltage from Time Domain Solver (t-solver) results.

2.1 Theory

In the following, electromagnetic devices which use time varying electromagnetic fields to alter the momentum of charged particle beams are referred to as kickers. An electrode structure or cavity resonator is constructed in a way, that by feeding electromagnetic power to the structure, a certain mode of field distribution is formed (Fig. 2.1). One Figure of Merit (FOM) of kickers is how much momentum change Δp

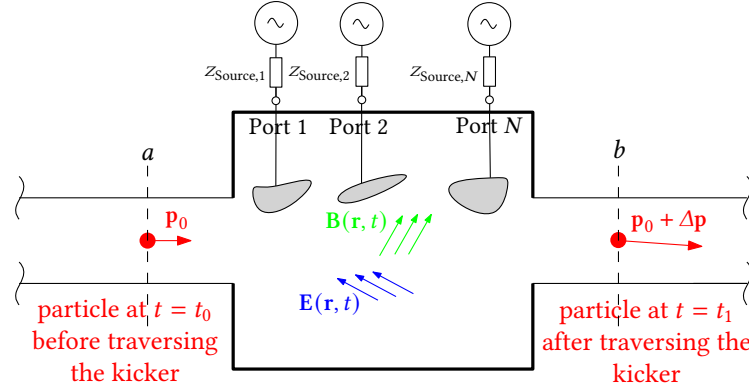


Fig. 2.1: Illustration of a kicker device, embedded into a beam pipe. The kicker may be powered by N ports. Depending on the feed signals, the electrode or cavity structure of the kicker forms a specific field distribution. E - and B -fields change the momentum of the charged particle traversing the kicker.

can be applied to a test particle with a given total power $P_{\text{total,inc}}$ incident to the kickers feeding ports. In this chapter, formulas for this FOM are derived and necessary approximations discussed. The concepts in this chapter follow the work by Goldberg and Lambertson [12]. The excited fields act on the particle with charge q and velocity $\mathbf{v}(t)$ via the Lorentz Force

$$\mathbf{F}(\mathbf{r}, t) = q(\mathbf{E}(\mathbf{r}, t) + \mathbf{v}(t) \times \mathbf{B}(\mathbf{r}, t)) . \quad (2.1)$$

Note that in the following, the fields stemming from the (moving) charged particle are neglected. As the velocity of charged particles in accelerator facilities approaches the speed of light, the relation of force acting upon the particle and its momentum needs to be considered in its relativistic form ¹

$$\mathbf{F}(\mathbf{r}, t) = \dot{\mathbf{p}}(\mathbf{r}, t) \quad (2.2)$$

$$\mathbf{p}(\mathbf{r}, t) = m_0 \gamma \dot{\mathbf{r}}(t) \quad (2.3)$$

$$\gamma = 1 / \sqrt{1 - (|\dot{\mathbf{r}}| / c)^2} . \quad (2.4)$$

¹Note that $\mathbf{v}(t) = \dot{\mathbf{r}}(t)$ and that m_0 is the rest mass of the test particle.

(2.2), (2.3) and (2.4) together with (2.1) can be used to derive the differential equation

$$m_0 \frac{d}{dt} \left(\frac{1}{\sqrt{1 - (|\dot{\mathbf{r}}(t)|/c)^2}} \cdot \dot{\mathbf{r}}(t) \right) = q \left(\mathbf{E}(\mathbf{r}, t) + \dot{\mathbf{r}}(t) \times \mathbf{B}(\mathbf{r}, t) \right). \quad (2.5)$$

The solution of (2.5) is the trajectory $\mathbf{r}(t)$ of the particle with initial conditions $\mathbf{r}(t_0)$, $\dot{\mathbf{r}}(t_0)$. After solving the equation for $\mathbf{r}(t)$, the momentum change applied to the charged particle by the kicker is defined by

$$\Delta \mathbf{p} = \mathbf{p}(t_1) - \mathbf{p}(t_0), \quad (2.6)$$

which is equivalent to integrating the force acting on the particle from t_0 to t_1

$$\Delta \mathbf{p} = \int_{t_0}^{t_1} \mathbf{F}(\mathbf{r}(t), t) dt. \quad (2.7)$$

(2.7) is the starting point for simplifying the computation of the momentum change. From this point onwards, z will be used as longitudinal coordinate and x, y will be the transverse coordinates.

The longitudinal velocity of a particle circulating in an accelerator ring is typically multiple orders of magnitude larger than the transverse velocity. If a small change of the longitudinal momentum of the particle is assumed, the trajectory of a particle traversing the kicker can be approximated to first order by a straight, uniform motion

$$\mathbf{r}(t) = (v_z (t - t_0) + a) \mathbf{e}_z. \quad (2.8)$$

(2.8) can then be used to approximately evaluate (2.7) for given electromagnetic fields inside the kicker. To this end

$$z(t) = v_z (t - t_0) + a \quad (2.9)$$

needs to be rearranged for t and inserted into (2.7). By substitution then follows

$$\Delta \mathbf{p}(x, y) = \frac{1}{v_z} \int_a^b \mathbf{F}(x, y, z, t(z)) dz. \quad (2.10)$$

Assuming time-harmonic fields $\mathbf{E}(\mathbf{r}, t) = \text{Re}\{\mathbf{E}(\mathbf{r}, \omega)e^{j\omega t}\}$ and $\mathbf{B}(\mathbf{r}, t) = \text{Re}\{\mathbf{B}(\mathbf{r}, \omega)e^{j\omega t}\}$ leads to the expression

$$\Delta \mathbf{p}(x, y, \omega, \beta) = \frac{q}{v} e^{j\omega(t_0 - a/v)} \int_a^b \left(\mathbf{E}(\mathbf{r}, \omega) + v \mathbf{e}_z \times \mathbf{B}(\mathbf{r}, \omega) \right) e^{j\frac{\omega}{v}z} dz \text{ with } v_z = v = \beta c. \quad (2.11)$$

Several facts are worth noting at this point. Firstly, the complex exponent in front of the integral in (2.11) may be omitted by assuming the particle is at position a at the time t_0 . Secondly, only the z -component of the E -field is responsible for longitudinal momentum change, while both transverse E - and transverse B -fields are responsible for transverse momentum change. The longitudinal component of the B -field neither changes the longitudinal, nor the transverse component of the momentum. Moreover it can be seen, that for time-harmonic fields the momentum change not only depends on frequency and the transverse coordinates, but also on the beam velocity v .

β is the beam velocity v over the speed of light c and is usually used instead of v as a variable of the momentum change $\Delta \mathbf{p}$. At the first stochastic cooling plateau of the AD (refer to Fig. 1.2), $\beta = 0.96724$ and at the second stochastic cooling plateau $\beta = 0.90532$. These two values are used throughout Chapter 3 for computing the momentum change.

From the momentum change, the so-called beam voltage $V(x, y, \omega)$ can be defined by

$$V(x, y, \omega, \beta) = \frac{v}{q} \Delta \mathbf{p}(x, y, \omega, \beta) = \int_a^b \left(\mathbf{E}(\mathbf{r}, \omega) + v \mathbf{e}_z \times \mathbf{B}(\mathbf{r}, \omega) \right) e^{j\frac{\omega}{\beta c}z} dz. \quad (2.12)$$

The momentum change and beam voltage can be decomposed into their longitudinal and transverse components. The notation

$$\Delta \mathbf{p} = \Delta \mathbf{p}_{\parallel} + \Delta \mathbf{p}_{\perp} \quad (2.13)$$

where $\Delta p_{\parallel}$ is the longitudinal and $\Delta p_{\perp}$ is the transverse component is used in what follows.

The shunt impedance $Z_{\text{shunt}}(x, y, \omega, \beta)$, is introduced as a FOM of kickers, having longitudinal and transverse components

$$Z_{\text{shunt}\parallel/\perp}(x, y, \omega, \beta) = \frac{|V_{\parallel/\perp}(x, y, \omega, \beta)|^2}{2P_{\text{total,inc}}} . \quad (2.14)$$

The shunt impedance is proportional to the absolute squared value of the momentum change of a test particle of charge q over the total power incident to the feeding ports of the kickers. Its physical dimension is Ω . Note, that for kicker designs with multiple feed ports, the shunt impedance depends on the phasing of the excitation ports with respect to each other. This fact is treated in Section 2.4.

A dimension-less FOM of kickers is the kick factor

$$\mathbf{K}(x, y, \omega, \beta) = \frac{\mathbf{V}(x, y, \omega, \beta)}{V_{\text{inc}}} \quad (2.15)$$

with $V_{\text{inc}} = \sqrt{2Z_0P_{\text{total,inc}}}$, Z_0 being the reference impedance. By the chosen definitions, shunt impedance and kick factor are related by

$$Z_{\text{shunt}\parallel/\perp}(x, y, \omega, \beta) = Z_0 |K_{\parallel/\perp}(x, y, \omega, \beta)|^2 . \quad (2.16)$$

Relation between longitudinal and transverse kick behavior

Using the previously assumed simplifications, Faraday's law and some vector calculus identities, the Panowsky Wenzel Theorem

$$\frac{\partial}{\partial t} \Delta p_{\perp} = -v \nabla_{\perp} (\Delta p_{\parallel}) + q(E_{\perp}(b, t_B) - E_{\perp}(a, t_A)) \quad (2.17)$$

was derived in [12] for arbitrarily time varying fields. The original article of Panowsky and Wenzel [13] contains a derivation for time-harmonic fields. Rearranging (2.17) leads to the formulation in terms of beam voltage

$$\frac{\partial}{\partial t} \frac{1}{v} \Delta V_{\perp} = -\nabla_{\perp} (\Delta V_{\parallel}) + E_{\perp}(b, t_B) - E_{\perp}(a, t_A) . \quad (2.18)$$

For time-harmonic fields, the time derivative on the left side is replaced by $j\omega$. Furthermore, the term including the transverse E -field components at the ends of the kicker may be omitted, if they are sufficiently small.² Using the Panowsky Wenzel Theorem, both longitudinal and transverse momentum changes can be computed, only requiring knowledge of the longitudinal component E_z of the E -field.

Relation between Kickers and Pickups

Pickups are components, used as sensors for the beam current I_B , the dipole moment $\Delta x I_B$ or higher moments of a beam with respect to the main beam axis. Fig. 2.2 shows an illustration of a pickup. Its FOM are the longitudinal transfer impedance

$$Z_{P\parallel} = \frac{V_{B\parallel}}{I_B} \quad (2.19)$$

and the transverse transfer impedance

$$Z_{P\perp} = \frac{V_{B\perp}}{\Delta x I_B} \quad (2.20)$$

where the voltages $V_{B\parallel}$ and $V_{B\perp}$ depend on the voltages at the N outputs of the pickup. A suitable combiner network is needed to obtain optimal performance. An example with a suitable combiner network is shown in Section 2.4. Every kicker can be used as a pickup and vice versa. Non-reciprocal devices are excluded

²For example, if the integration limits a, b are set far away from the kicker structure and the structure is excited below the lowest cut-off frequency of the beam pipes.

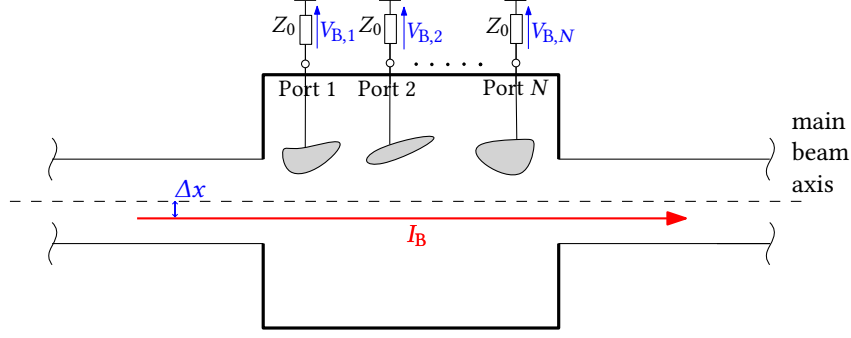


Fig. 2.2: Illustration of an N -port pickup device. The output signals depend on the total beam current I_B and the dipole moment $\Delta x I_B$ or higher moments of a beam with respect to the main beam axis.

here. The pickup behavior (*i.e.*, power of output signal in dependence of the beam current) is coupled to the devices behavior as a kicker via the Reciprocity Theorem of Electromagnetics. Assuming a single TEM signal port and no reflections at the signal port, [12] derived from the Reciprocity Theorem that the longitudinal FOM's of pickups and kickers are related by

$$Z_{P\parallel} = \frac{Z_0}{2} K_{\parallel}. \quad (2.21)$$

Using additionally the Panowsky Wenzel Theorem,

$$Z_{P\perp} = -j\omega \frac{Z_0}{2\beta c} K_{\perp} \quad (2.22)$$

was derived in [12] with the important side note: “For electrode systems which exhibit directional behavior (*e.g.*, striplines) the direction in which the beam passes when the device operates as a pickup must be opposite that which it does when the device acts as a kicker”[12, p. 20]. This fact is used in Section 2.2 and Appendices A to C for derivations of kick factor and shunt impedance of certain kicker structures.

The behavior (pickup or kicker) which is more simple to analyze can be chosen for direct analysis (numerical field simulation or analytical approximations) and the other one can be calculated from the formulas (2.21) and (2.22). For example, the pickup behavior of the stripline topology will be analytically analyzed in Section 2.2. Its kick behavior will then be calculated from the equations (2.21) and (2.22).

2.2 Stripline Topology

Broadband pickups and kickers with approx. octave bandwidth can be realized in the stripline topology. Stripline pickups and kickers consist of conductive strips which form TEM lines together with the beam pipe or other conductive parts inside the beam pipe.

In this section, qualitative explanations of the working principles of stripline kickers and pickups are given and analytic approximations for the transfer impedance, kick factor and shunt impedance are provided. The devices used in the AD stochastic cooling pickups and kickers are of a modified stripline topology. This topology and its generalization are treated in Section 2.2.4.

2.2.1 The Pickup Working Principle

An illustration of a stripline pickup with a 180° hybrid connected to its ports is shown in Fig. 2.3. Assume a charged particle traversing the pickup at relativistic velocity. When the charged particle enters the pickup, so will its image on the beampipe walls. The image current transitions from the beampipe onto the striplines as a displacement current which excites voltage pulses (waves) on the striplines.

The backward-travelling pulses appears at the terminals A and B of the stripline electrodes and a forward-travelling wave propagates along the striplines. The forward travelling pulses are then absorbed in the terminations of the striplines. When the charged particle exits the striplines, its image charge has to transition from the striplines to the beampipe walls as a displacement current, exciting voltage pulses of negative polarity on the striplines. The backward traveling wave exits the terminals A and B of the striplines at a time delay of $L(1/v_{\text{beam}} + 1/v_{\text{line}})$ with respect to the first voltage pulses, whereas the forward travelling pulses are again absorbed in the terminations of the striplines.

The impulse response of a stripline pickup is thus two impulses of opposite polarity with a time-delay of $L(1/v_{\text{beam}} + 1/v_{\text{line}})$ in between. This leads to sine-wave-like frequency response. The output voltage V_Σ of the hybrid contains information on the beam current I_b and the output voltage V_Δ of the hybrid contains information on the dipole moment $\Delta x I_b$ of the beam. Detailed analytical treatment of the pickup behavior can be found in [12], [14]–[16] and in Appendix A.

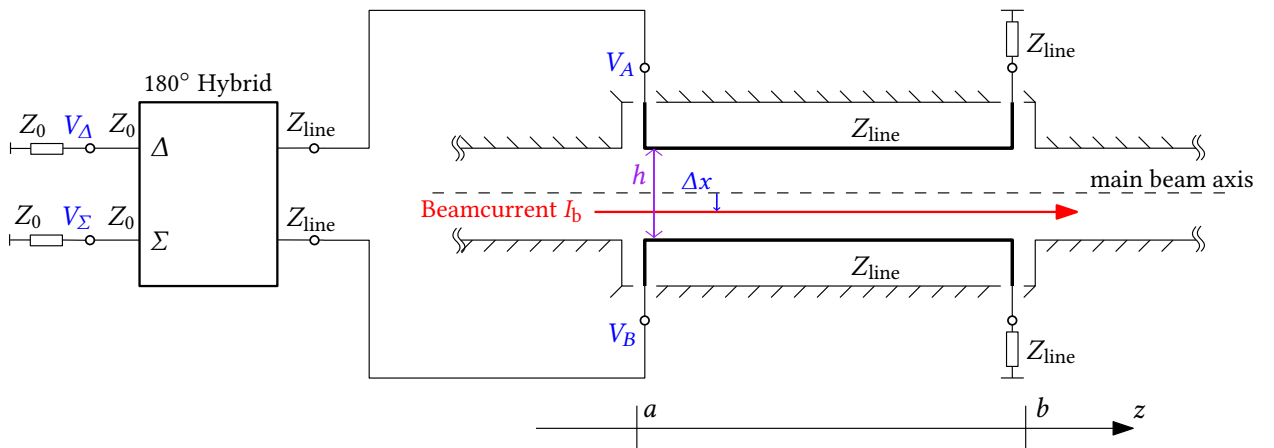


Fig. 2.3: Illustration of a stripline pickup with a 180° hybrid. The length of the striplines is $L = b - a$.

2.2.2 The Kicker Working Principle

For analyzing the longitudinal kick behavior to a positively charged test particle, only the z -component of the E -field in the kicker needs to be observed. Since the stripline is a TEM waveguide, only the regions at the ends of the stripline, where end effects lead to a non-zero z -component of the E -field are of interest.

Note the direction of the z -component of the E -field in Fig. 2.4. This chosen convention leads to the E -field vector at $z = a$ pointing in positive z -direction, if the voltage V_{line} on the line at $z = a$ is negative. By the same argument, the E -field vector at $z = b$ points in positive z -direction, if the voltage V_{line} on the line at $z = b$ is positive. From this follows, that the longitudinal momentum of the test particle cannot be changed when driving the stripline at DC, as the particle will experience the same absolute value of force at the two ends of the stripline, but always at different sign.

Let the line now be excited at the downstream end³ by a sinusoidal voltage. Furthermore, it is assumed that the section at the ends of the striplines, where interaction of E_z and the beam happens, is infinitely short. It is assumed, that the imposed momentum change will be small and that the particle velocity v_{beam} is equal to the phase velocity v_{line} of the line. It then follows, that the timing of the peak positive line voltage V_{line} and peak negative line voltage values align correctly for maximum positive momentum change of the test particle, if the line is driven at the frequency, where it is a quarter wavelength long (refer also to the right side of Fig. 2.4).

At other frequencies, the longitudinal momentum change imposed on the test particle will be smaller. This behavior periodically continues with peaks at odd integer multiples of the frequency where the striplines are a quarter wavelength long. It is crucial that the feed direction is opposite to the direction of the beams trajectory, as otherwise the kicks at the ends of the stripline would cancel each other.⁴

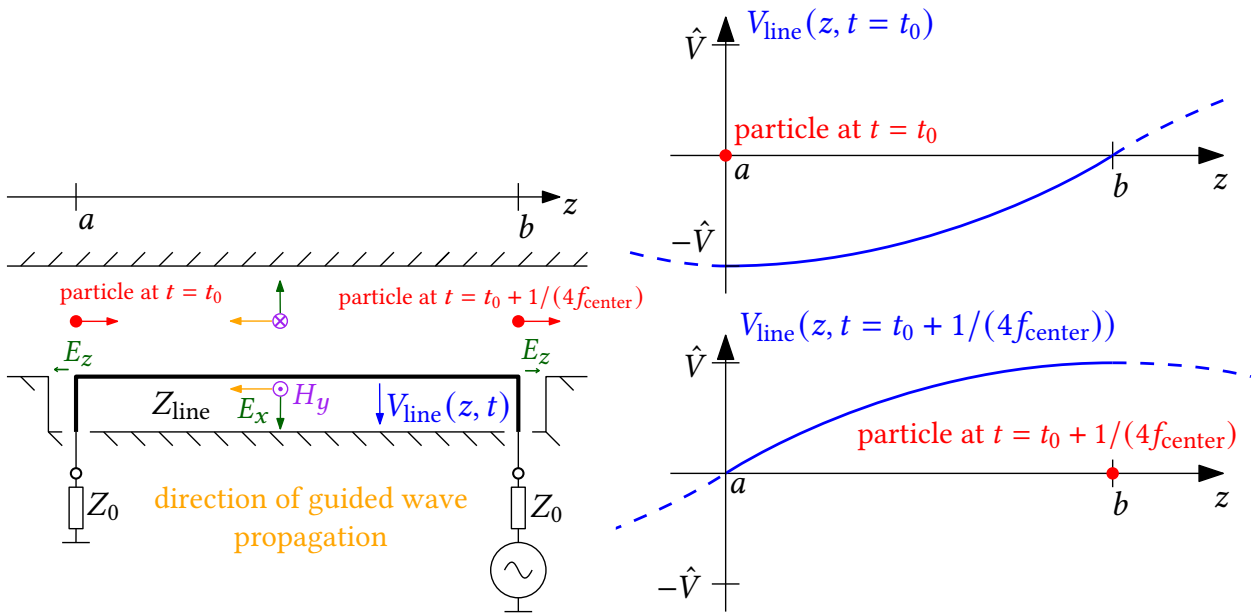


Fig. 2.4: Illustration of the interaction of a charged particle with the longitudinal E -field created by a single stripline. The stripline is driven at the frequency f_{center} , where its length is equal to $\lambda_{\text{center}}/4$. $v_{\text{beam}} = v_{\text{line}}$ and $Z_{\text{line}} = Z_0$ are assumed.

By analogous reasoning, the transverse kick behavior can be qualitatively described. Note that both transverse E - and B -fields contribute to the transverse kick to the test particle and that the fields interact with the beam over the full length of the line. In contrast to the longitudinal behavior, the maximum transverse kick is achieved at DC. This is due to the fact, that only at DC the transverse E - and B -fields are the most “uniform” and always in the correct direction across the complete length of the line. For non-zero frequencies, the transverse momentum change will decrease. The stripline thus has low-pass kick characteristics.

³Upstream is the end of the stripline where the particle beam is coming from, downstream is the end of the stripline where the particle beam is travelling to.

⁴For $v_{\text{beam}} \neq v_{\text{line}}$, kicks are possible if the feed and beam directions are equal.

Using two striplines, mounted to opposite sides of the beam pipe is advantageous. The longitudinal kick solely depends on the common mode excitation voltage and the transverse kick solely depends on the differential mode excitation voltage of the lines. Therefore, one device can be used to independently change the transverse and longitudinal momentum of a beam traversing it. A 180° Hybrid may be used to independently excite the two lines in common and differential mode by the use of two drive signals, V_Δ and V_Σ (refer to Fig. 2.5).

The fact that for in-phase excitation of the striplines, no transverse kick is given to a charged particle traversing the kicker through the center of the beam pipe gets clear when observing the transverse field distribution for such an excitation (Fig. 2.6). It can be seen, that the transverse E- and transverse B-fields are zero in the center of the beam pipe. The transverse fields for 180° out of phase excitation of the striplines can be seen in Fig. 2.7.

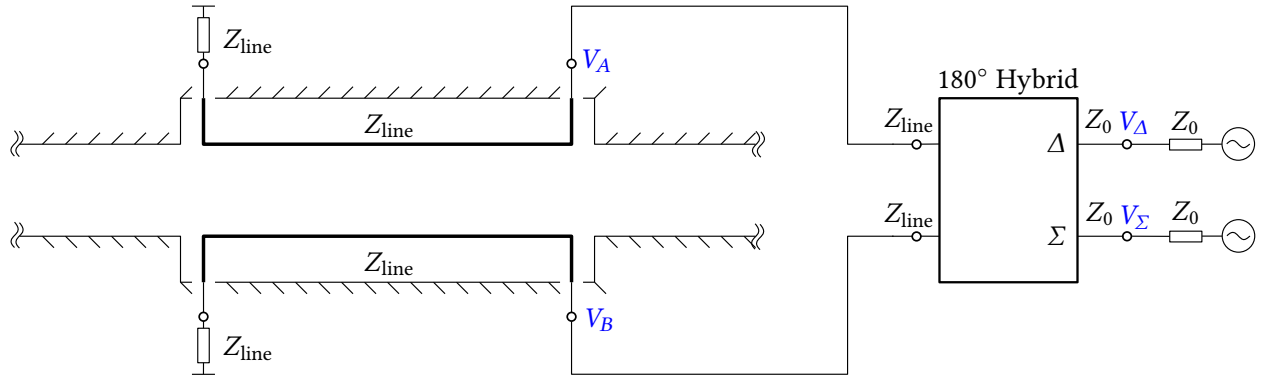


Fig. 2.5: Illustration of a stripline kicker with hybrid.

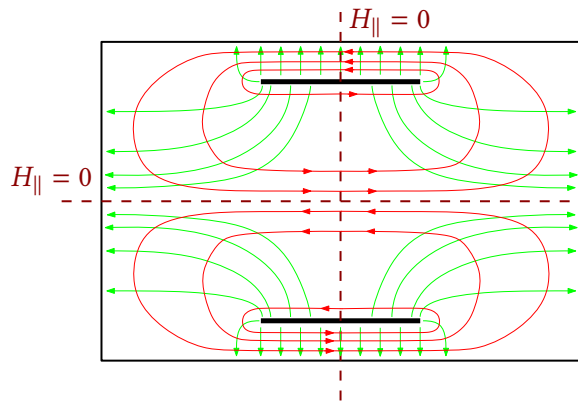


Fig. 2.6: Field symmetry, Σ -mode excitation.

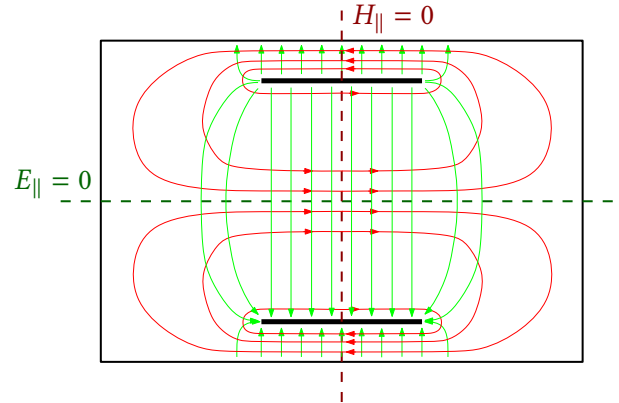


Fig. 2.7: Field symmetry, Δ -mode excitation.

2.2.3 Analytical Approximations

Analytical approximations of the transfer impedance are provided in Tab. 2.1 and analytical approximations of the kick factor and shunt impedance are given in Tab. 2.2. Derivations of the analytical formulas can be found in [12]. More detailed derivations are provided in Appendix A. The longitudinal and transverse geometry factors

$$g_{\parallel} = \frac{2}{\pi} \arctan \left(\sinh \left(\frac{\pi w}{2h} \right) \right) [12, (8.8)] \quad (2.23)$$

$$g_{\perp} = \tanh \left(\frac{\pi w}{2h} \right) [12, (8.13)] \quad (2.24)$$

which are used in the formulas, depend on the width w of the striplines and the distance h between the striplines. L is the length of the striplines, Z_{line} is the line impedance of the striplines and Z_0 is the reference impedance. The beam velocity is v_{beam} and the phase velocity of the striplines is v_{line} .

Tab. 2.1: Summary of the pickup behavior of striplines. Note that $\theta = \omega L(1/v_{\text{beam}} + 1/v_{\text{line}})/2$. The formulas are equal to (8.10) and (8.16) in [12].

	Transfer impedance Z_p
longitudinal	$\sqrt{\frac{Z_{\text{line}} Z_0}{2}} g_{\parallel} \sin(\theta) \exp(j(\pi/2 - \theta))$
transverse	$\sqrt{\frac{Z_{\text{line}} Z_0}{2}} g_{\perp} \frac{2}{h} \sin(\theta) \exp(j(\pi/2 - \theta))$

Tab. 2.2: Summary of the kick behavior of striplines. Note that $\theta = \omega L(1/v_{\text{beam}} + 1/v_{\text{line}})/2$. The kick factor formulas are equal to (8.9) and (8.15a) in [12] and shunt impedance formulas are equal to (8.11) and (8.17) in [12].

	Shunt impedance Z_{shunt}	Kick factor K
longitudinal	$2Z_{\text{line}} g_{\parallel}^2 \sin^2(\theta)$	$2\sqrt{\frac{Z_{\text{line}}}{2Z_0}} g_{\parallel} \sin(\theta) \exp(j(\pi/2 - \theta))$
transverse	$2Z_{\text{line}} g_{\perp}^2 \left(\frac{L}{h}\right)^2 \left(1 + \frac{v_{\text{beam}}}{v_{\text{line}}}\right)^2 \text{sinc}^2(\theta)$	$2\sqrt{\frac{Z_{\text{line}}}{2Z_0}} g_{\perp} \frac{L}{h} \left(1 + \frac{v_{\text{beam}}}{v_{\text{line}}}\right) \text{sinc}(\theta) \exp(-j\theta)$

Several facts merit attention concerning the results (refer to Fig. 2.8). First of all, both pickup and kicker responses have larger transverse peak values than longitudinal peak values. Secondly, the longitudinal and transverse transfer impedances both have the same periodic behavior over frequency, with band-pass behavior around odd integer multiples of $\theta = \pi/2$. Furthermore, while the longitudinal kick behavior has the same shape over frequency, the transverse kick behavior has sinc-dependence over frequency.

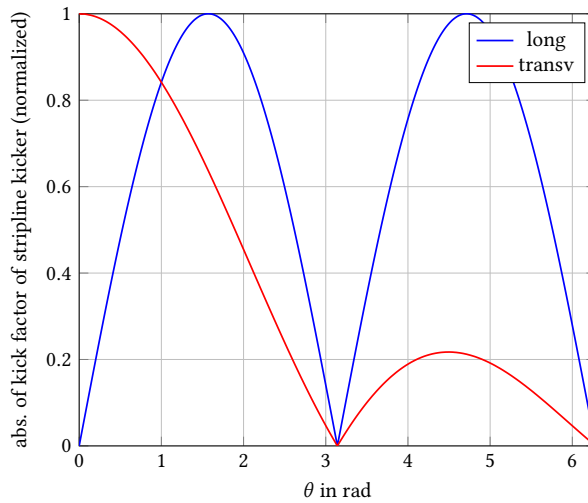


Fig. 2.8: Long. and transv. kick factor of a stripline kicker, normalized to their peak values each.

2.2.4 Multi-Loops

A gain in peak value of the transfer impedance and kick factor compared to the stripline responses in Section 2.2.3 can be obtained by using a modified stripline design (Fig. 2.9). This design is called multi-loop. In the course of this work, classic striplines are also referred to as single-loops and double-loops are also referred to as superelectrodes. If M striplines are cascaded by $\lambda/2$ delay-lines, a gain of $\sqrt{M}$ is achieved in transfer impedance and kick factor, while the 3 dB-bandwidth is approximately reduced by a factor of $0.9/M$ compared to a classic stripline [17, p. 398].

In the AD stochastic cooling pickups and kickers, double-loops are used (also referred to as super-electrodes) [11], [17], [18]. In double-loops, two $\lambda/4$ long striplines are connected by a $\lambda/2$ long delay-line.

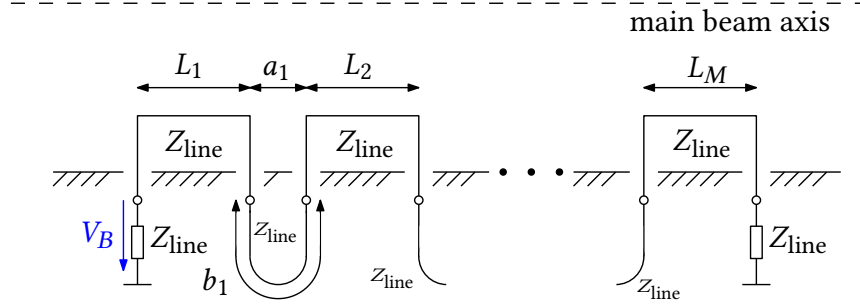


Fig. 2.9: Illustration of the lower half of an M element multi-loop for general dimensions.

As derived in Appendix B, the response of a multi-loop can be computed by multiplying the intra-loop array factor to the response function of a single stripline

$$Z_{P,\text{multi-loop},\parallel/\perp} = Z_{P,\text{single-loop},\parallel/\perp} \cdot \text{AF}_{\text{intra-loop}} \quad (2.25)$$

$$K_{\text{multi-loop},\parallel/\perp} = K_{\text{single-loop},\parallel/\perp} \cdot \text{AF}_{\text{intra-loop}} \quad (2.26)$$

$$Z_{\text{shunt},\text{multi-loop},\parallel/\perp} = Z_{\text{shunt},\text{single-loop},\parallel/\perp} \cdot |\text{AF}_{\text{intra-loop}}|^2. \quad (2.27)$$

The intra-loop array factor AF is the same for the longitudinal and transverse responses of transfer impedance and kick factor. For the shunt impedance, the absolute squared value has to be used.

For a multi-loop device of M stripline elements of equal length L , delay-lines of equal length b and constant spacing a between the striplines, the intra-loop array factor is

$$\text{AF}_{\text{intra-loop}} = \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)} \exp\left(-j[(M-1)(\theta + \zeta)]\right) \quad (2.28)$$

with

$$\theta = \frac{\omega}{2} L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) \quad (2.29)$$

$$\zeta = \frac{\omega}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right). \quad (2.30)$$

A comparison of the total shunt impedance of a double-loop and two independently fed single-loops is shown in Figs. 2.10 and 2.11. It can be seen, that longitudinally there is a 3 dB improvement in peak value of the double-loop and a reduction in bandwidth, compared to the arrangement of two independently fed single-loops. Furthermore, the double-loop has a second pass-band in the transverse response around the design-frequency compared to the single-loop.

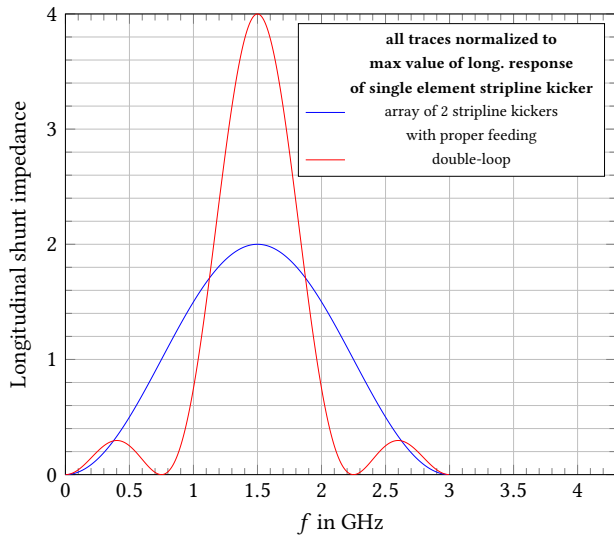


Fig. 2.10: Comparison of long. shunt impedance of single- & double-loop.

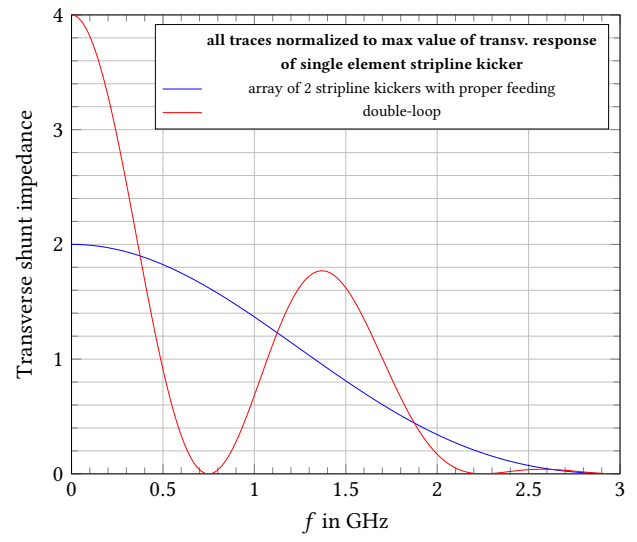


Fig. 2.11: Comparison of transv. shunt impedance of single- & double-loop.

2.3 Slotline Topology

The slotline topology was initially proposed by Caspers in 1987 [19]. This topology offers higher gain-bandwidth product over unit-length than striplines and a more homogeneous field distribution on the beam aperture. The basic setup of this topology consists of a slot in a ground-plane, which interacts with the charged particle beam traversing the slot. The slot is coupled to by a transmission line. This is illustrated by the dashed lines in Figs. 2.12 and 2.13.

Two subcategories were introduced by Caspers [19]: The Transverse Standing Wave Device sub-group of slotline devices (TSWS) (Fig. 2.12) and Transverse Travelling Wave Device sub-group of slotline devices (TTWS) (Fig. 2.13). TTWS are terminated by a load, while resonant devices are not. Very high relative bandwidths (up to a decade according to [19]) can be achieved with the TTWS. If a relative bandwidth in the octave range is of interest, the TSWS is better suited as it allows for a larger peak value than the TTWS.

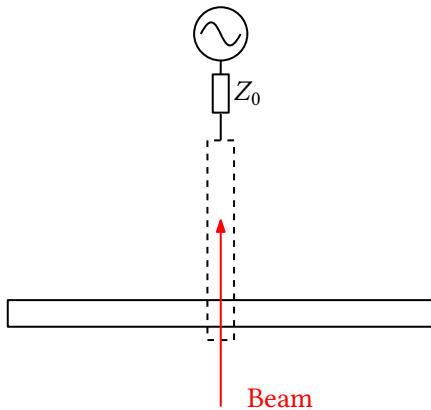


Fig. 2.12: Illustration of a resonant slotline kicker (TSWS).

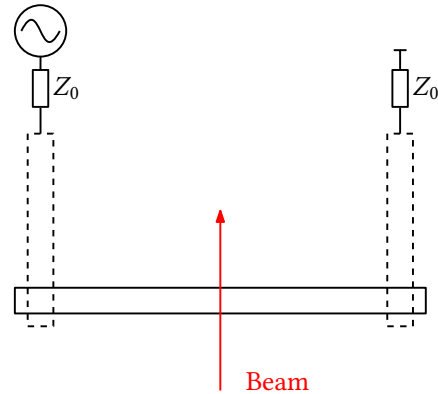


Fig. 2.13: Illustration of a travelling wave slotline (TTWS) kicker like it was investigated in [20].

In kicker operation, both subcategories work similar to striplines. Transverse and longitudinal kicks are imposed on a charged particle beam by driving slots on the two sides of the beam either in phase or 180° out of phase. An array effect is achieved by driving all slots (refer to Fig. 2.14) with a time delay scheme that matches the beam's time of flight through the array (refer to Section 2.4).

In pickup operation, the image current of the beam induces traveling waves on the slotline, as the image current traverses the slot as a displacement current. These wave can be coupled out of the slot to a transmission line. There are several schemes of how to couple a transmission line to a slot. Furthermore, two transmission lines can be coupled to the slot as has shown in [20]–[26], improving the sensitivity in terms of output power by a factor of 2.

In [20], results of CST simulations of the TTWS and stripline topology have been compared. It can be seen, that much flatter frequency responses can be achieved with the TTWS compared to stripline devices. It was also shown, that the kick factor and transfer impedance are less sensitive to the transverse position of the beam.

The dimensions of unit cells of kickers and pickups scale inversely proportional to the center frequency of the used frequency band. Above the single digit GHz-range, all dimensions of striplines are in the single digit mm-range to μm -range. These dimensions are hard to machine at low cost and high precision. Slotlines on the other hand can be constructed by techniques of Printed Circuit Board (PCB)-manufacturing which allows tolerances down to the μm -range. There is one variant of the stripline type, which may also be constructed by means of PCB-manufacturing: Planar loops. This variant is not subject of this work and it is referred to [15], [27]–[29] for theory, design and measurements of such printed loops.

Another slot-type device consists of an array of slots, where all slots on one side of the beam are coupled to one TEM-line instead of using power combiners to combine the signals from all slots [30]–[32]. This topology is not considered in this work, as it is not suitable for operation at different beam velocities.

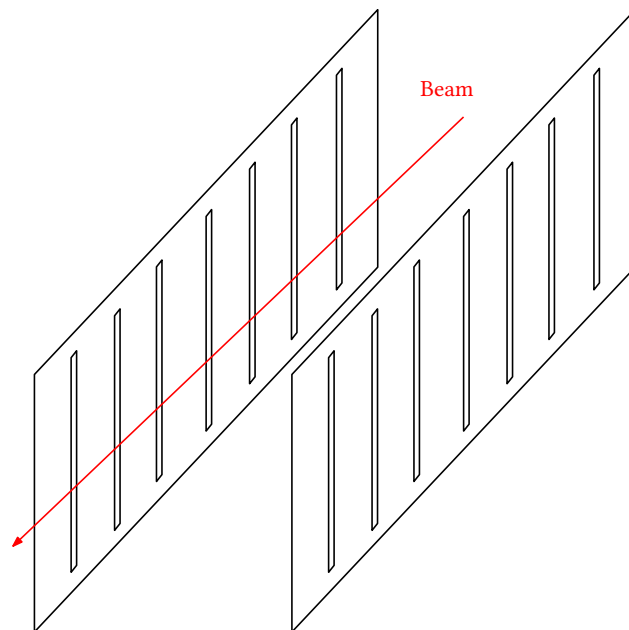


Fig. 2.14: 3D illustration of a slotline kicker/pickup array.

2.4 Pickup and Kicker Arrays

Broadband high sensitivity pickups and broadband high strength kickers can be realized by using single pickup/ kicker elements together with a suitable combiner or divider network. In this section, the working principle of pickup and kicker arrays is explained in time- and frequency domain. Furthermore, the array factor of such arrays is derived and discussed.

Assume an array of N ideal pickups⁵ that are connected to an ideal power combiner network with equal time delay (refer to Fig. 2.15). At the outputs of each pickup, a pulse is launched into the power combiner

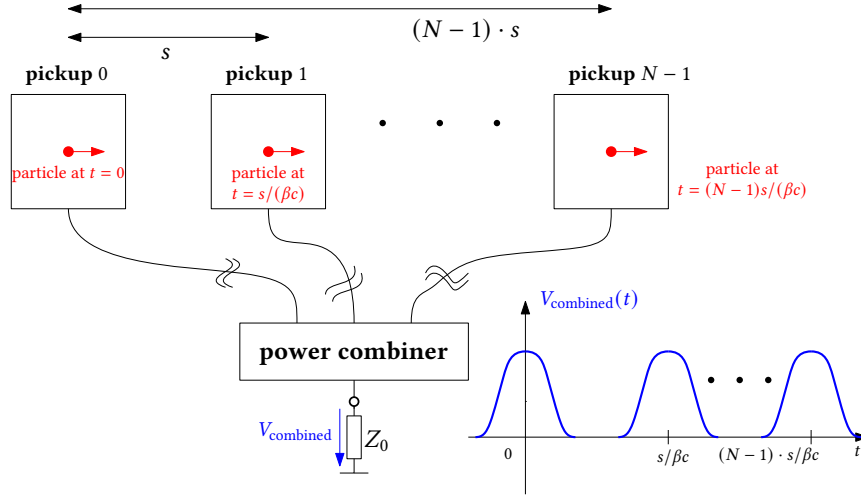


Fig. 2.15: A pickup array. The lines from the pickups to the power combiner are of equal length.

when a charged particle traverses a pickup. As the time delay from the pickups to the power combiner is equal, the output voltage of the combiner consists of a train of N delayed pulses

$$V_{\text{combined}}(t) = \frac{1}{\sqrt{N}} \cdot \sum_{m=0}^{N-1} V_{\text{pickup}}(t - m(s/\beta c)). \quad (2.31)$$

In frequency domain this gives the response

$$V_{\text{combined}}(f) = V_{\text{pickup}}(f) \cdot \frac{1}{\sqrt{N}} \cdot \sum_{m=0}^{N-1} \exp(-j\omega m(s/\beta c)) \quad (2.32)$$

$$V_{\text{combined}}(f) = V_{\text{pickup}}(f) \cdot \frac{1}{\sqrt{N}} \cdot \frac{\sin\left(\frac{N\omega s}{2\beta c}\right)}{\sin\left(\frac{\omega s}{2\beta c}\right)} \cdot \exp\left(-j\omega \frac{s}{2\beta c}(N-1)\right). \quad (2.33)$$

The array thus gives a DC gain of $\sqrt{N}$ in the output voltage compared to a single pickup element, but severely limits the bandwidth. The total transfer impedance is then

$$Z_{\text{pickup,total}} = \underbrace{\frac{1}{\sqrt{N}} \cdot \frac{\sin\left(\frac{N\omega s}{2\beta c}\right)}{\sin\left(\frac{\omega s}{2\beta c}\right)} \cdot \exp\left(-j\omega \frac{s}{2\beta c}(N-1)\right)}_{\text{Array Factor AF}} \cdot Z_{\text{pickup,single}}. \quad (2.34)$$

An array architecture of improved bandwidth performance is created by making all the voltage pulses overlap in time-domain, by accounting for the time-of-flight of the beam by inserting delay-lines between the pickup elements and the input ports of the power combiner (refer to Fig. 2.16). This architecture gives a broadband array factor of $\sqrt{N}$ for the transfer impedance, if the time of flight of the beam is matched by the delay lines. In that case, only the characteristics of the used power combiner limit the bandwidth.

⁵Ideal pickups have unity gain, no distortion, and are matched to the reference impedance.

The time-domain output voltage of the power combiner network is then

$$V_{\text{combined}}(t) = \frac{1}{\sqrt{N}} \cdot \sum_{m=0}^{N-1} V_{\text{pickup}}(t - m(s/\beta c - t_m)) \quad (2.35)$$

and the corresponding frequency domain response is

$$V_{\text{combined}}(f) = V_{\text{pickup}}(f) \cdot \frac{1}{\sqrt{N}} \cdot \sum_{m=0}^{N-1} \exp(-j\omega m(s/\beta c - t_m)) \quad (2.36)$$

with t_m being the delays provided by the transmission lines from the pickups to the combiner network (refer to Fig. 2.16). The total transfer impedance for arbitrary but equal time delays $t_m = \Delta t$ in the combiner network is

$$Z_{\text{pickup,total}} = \underbrace{\frac{1}{\sqrt{N}} \cdot \frac{\sin\left(\frac{N\omega}{2}\left(\frac{s}{\beta c} - \Delta t\right)\right)}{\sin\left(\frac{\omega}{2}\left(\frac{s}{\beta c} - \Delta t\right)\right)} \cdot \exp\left(-j\frac{(N-1)\omega}{2}\left(\frac{s}{\beta c} - \Delta t\right)\right)}_{\text{Array Factor AF}} \cdot Z_{\text{pickup,single}} \quad (2.37)$$

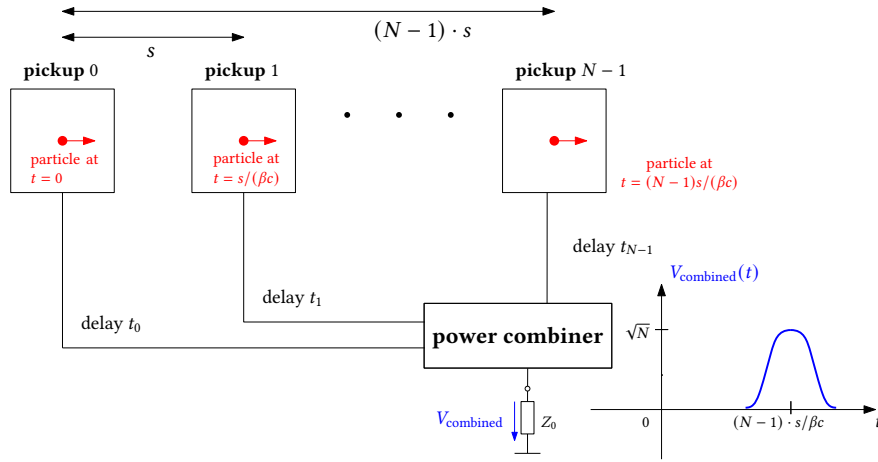


Fig. 2.16: A pickup array. The lines from the pickups to the power combiner account for the time of flight of the beam.

Fig. 2.17 shows the absolute value of array factors of 24 element arrays with $s = 87.2$ mm spacing between the pickup elements⁶. It can be seen, that the bandwidth of the array factor reduces when the delays in the combiner network do not match the beam time of flight up to the most extreme case of no time of flight compensation, where the bandwidth is severely reduced. For systems which operate at different beam velocities such as the stochastic cooling system of the AD, switchable delays may be used to preserve comparable sensitivity and bandwidth performance. At the moment, this technique is not used for the stochastic cooling pickups. All delays for the pickups are matched to $\beta = 0.96724$. This leads to an array factor of the pickups at $\beta = 0.90532$, seen in the green trace of Fig. 2.17.

The functionality of kicker arrays can be discussed analogously to the theory of pickup arrays. In time-domain, the "array effect" is achieved by having lots of individual kick elements being driven by delayed electromagnetic pulses, such that over the full length of the array the kicks of the single elements add up coherently. In contrast to the pickup array from Fig. 2.16, the combiner is now used as a power divider and the delays of the transmission lines from the power divider to the kick elements have to be arranged in opposite order for correct timing of the kicks to the particle traversing the kicker array. If the delays of pickup and kicker arrays are as described, the array factor of the pickup array is equal to the array factor of the kicker array due to reciprocity of Maxwell's equations.

⁶This spacing is used in the AD stochastic cooling pickups and kickers (refer to Fig. 3.2).

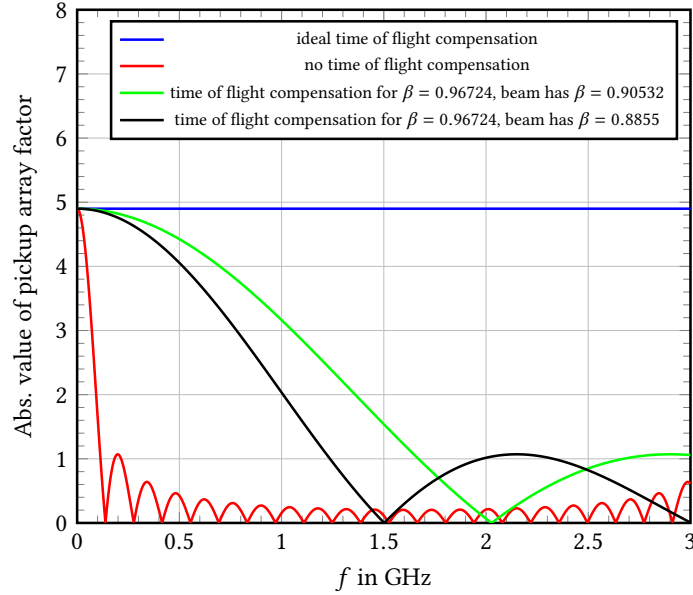


Fig. 2.17: Comparison of array factors of a 24 element pickup array. The element spacing is $s = 87.2$ mm.

To conclude, the response of a pickup or kicker array can be computed by multiplying the response of a single element by the array factor AF of the array (refer to (2.38) and (2.39)). This is analogous to antenna arrays, where the far-field radiation pattern of an array can be computed by multiplying the radiation pattern of a single antenna by the array factor of the array.

$$Z_{P,\parallel/\perp,\text{total}}(f, \beta) = Z_{P,\parallel/\perp,\text{single element}}(f, \beta) \cdot AF(f, \beta) \quad (2.38)$$

$$K_{\parallel/\perp,\text{total}}(f, \beta) = K_{\parallel/\perp,\text{single element}}(f, \beta) \cdot AF(f, \beta) \quad (2.39)$$

From (2.39) follows

$$Z_{\text{shunt},\parallel/\perp,\text{total}}(f, \beta) = Z_{\text{shunt},\parallel/\perp,\text{single element}}(f, \beta) \cdot |AF(f, \beta)|^2 \quad (2.40)$$

for the total shunt impedance of kicker array.

In practice, instead of one single power combiner for a pickup array or a single power splitter for a kicker array, multiple combiners/splitters are used. Keeping in mind, that the kicker/pickup array is contained inside a vacuum tank, a trade-off has to be made between having a low number of RF vacuum feedthroughs and having an array that can operate at different beam momenta with constant kick strength/ sensitivity. The only way of getting around this, is to have switchable delays inside the vacuum tank. Excluding the possibility of having switchable delay-lines inside the vacuum tank, this trade-off is explained in the following.

In one extreme case, every kick element is granted own RF vacuum feedthroughs. In this case, having switchable delays is a viable option and uncompromised performance is achieved at different beam momenta. Suppose having only one or two⁷ RF vacuum feedthroughs is desired and this while having a broadband high kick factor/ high transfer impedance device. In this case, all the power combination/ splitting has to be done inside the vacuum tank. In this case, the delays of the array are fixed and thus operation at beam velocities different from the design velocity leads to deteriorated array performance (refer to Fig. 2.17).

As mentioned before, the two pickup arrays used in the AD stochastic cooling system, have two vacuum feedthroughs each for the pickup signals. All the power combination from the 24 individual pickups is done inside the vacuum tank. The two AD stochastic cooling kickers on the other hand have a different scheme for driving all the individual kick elements. In the power splitting scheme of the kicker arrays, some splitting is done inside the vacuum tank, and some splitting is done outside of the vacuum tank with delays which switch to different values on the different stochastic cooling plateaus. In Appendix D, the array factor of the AD stochastic cooling kickers is derived and the power splitting scheme illustrated (refer to Fig. D.1).

⁷At least two RF feedthroughs are necessary, if transverse kicks/measurements are to be obtained by the array.

2.5 Performance Considerations on Phase Response and Bandwidth

This section addresses the phase response and bandwidth of kickers and pickups, building on the previously introduced theory of broadband pickup and kicker arrays.

The phase response of a kicker is the phase of the kick factor (which is equal to the phase of the beam voltage), while the phase of a pickup is the phase of the transfer impedance. As usual in linear systems which deal with the transmission of pulses, the unwrapped phase should be as linear as possible in the operational bandwidth of the system. A measure of the linearity of the phase response of a system is the nonlinear deviation from linear phase $\varphi_{\text{nonlinear}}$. The nonlinear phase is defined as the difference of the unwrapped phase to a linear fit to this unwrapped phase in the operational bandwidth

$$\varphi_{\text{nonlinear}}(f) = \varphi(f) - \varphi_{\text{linear}}(f). \quad (2.41)$$

In this work, the linear least-squares fit is used. A graphical illustration of the nonlinear phase deviation is shown in Fig. 2.18. The linear part of the phase only leads to a time delay of a modulated pulse propagating through the system, whereas higher order terms that are included in the nonlinear part lead to distortions. It is thus desired to keep the nonlinear deviation of the phase as small as possible in the operational bandwidth.

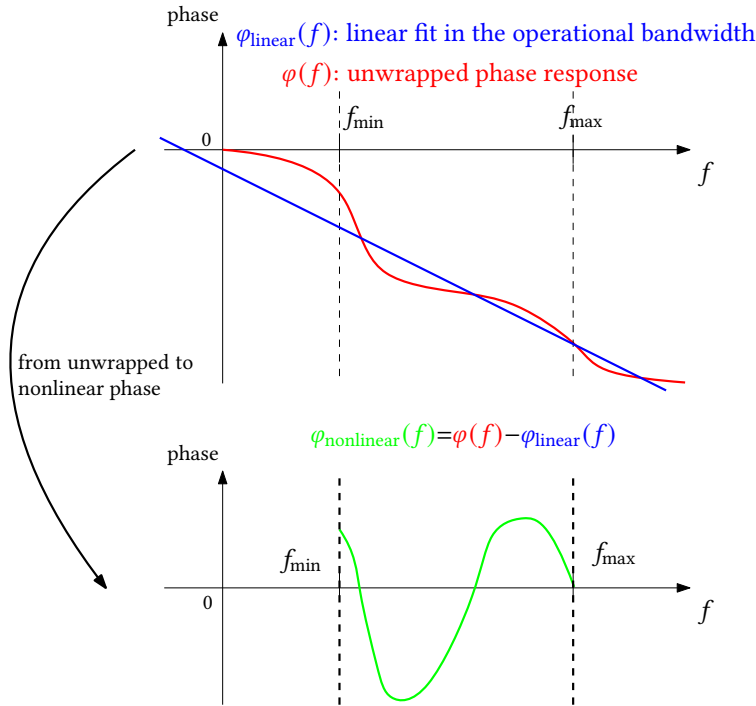


Fig. 2.18: Illustration of nonlinear phase.

Regarding bandwidth of pickups and kickers, there are two common definitions. The 3 dB bandwidth is defined by

$$\text{BW}_{3\text{dB}} = f_2 - f_1 \quad (2.42)$$

$$\text{with } f_{1,2} = \frac{1}{\sqrt{2}} \max(|K|). \quad (2.43)$$

A different definition involves integrating the absolute value of the kick factor or the shunt impedance over a certain frequency band and normalizing the result by the maximum value in this band:

$$\text{BW}_K = \frac{\int_{f_{\text{min}}}^{f_{\text{max}}} |K| df}{\max(|K|)} \quad (2.44)$$

$$\text{BW}_Z = \frac{\int_{f_{\text{min}}}^{f_{\text{max}}} |Z| df}{\max(|Z|)} \quad (2.45)$$

This definition is used in this work, as broadband kickers usually do not have steep roll-off behavior compared to RF-filters and amplifiers where the 3 dB definition is usually used. Both definitions may be applied to pickups, by replacing the kick factor by the transfer impedance.

To benchmark different pickup and kicker topologies, it is best to use the gain-bandwidth product over the length L of the kick element as figure of merit. Using this definition with the shunt impedance yields

$$\text{GBWP}/L = \frac{\max(Z_{shunt}) \cdot \text{BW}_Z}{L}. \quad (2.46)$$

When working with kick factors instead of shunt impedance,

$$\text{GBWP}/\sqrt{L} = \frac{\max(|K|) \cdot \text{BW}_K}{\sqrt{L}} \quad (2.47)$$

has to be used as figure of merit. This is due to the total kick factor of a properly phased array, scaling with the square root of number of unit cells instead of scaling linearly like the shunt impedance. The definition of the gain-bandwidth product over length including the kick factor has been used in [22], [23] for the optimization of slotline electrodes.

2.6 Aspects on Numerical Computation of Kickers

Typically, the performance of pickups is determined by means of electromagnetic simulations where the device is excited by a charged particle beam. Another way of numerically determining the response is to simulate the pickup in kicker operation and to compute the pickup response using the relations derived from the reciprocity theorem. In kicker operation, a structure is excited with electromagnetic power through one or several waveguide ports. After determining the complex electromagnetic fields inside the kicker structure, the integral in (2.11) has to be evaluated numerically for given values of β .

In the following, a few aspects on the numerical computation of kickers and thus also pickups are discussed. Both general and CST specific aspects are subject of the discussion.

2.6.1 Using Symmetry Planes in Simulations

Due to the symmetry of the kicker structures subject of this work, symmetry planes may be used to reduce the volume of the solution domain in numerical simulations while still obtaining correct results. Fig. 2.6 and Fig. 2.7 show the symmetry of a stripline kicker for common mode excitation (for longitudinal kicks) and differential mode excitation (for transverse kicks). Slotted stripline kickers also have this type of symmetry. Due to these symmetry planes, the kicker figures of merit of stripline and slotline structures can usually be obtained by electromagnetic simulations of a quarter of their total volume.

To numerically check this statement, a basic stripline kicker designed for a center frequency of 1 GHz (refer to Fig. 2.19) was simulated in CST for different permutations of symmetry planes. The simulated longitudinal and transverse kick factor data is shown in Fig. 2.20 to Fig. 2.23, both in absolute value and in phase. In each plot, the solid trace shows a result from a model without symmetry planes being used. The dashed traces show the difference of results of models that use symmetry planes with respect to the model that does not use symmetry planes. The legends of the plots are to be understood as follows: One symmetry plane means the horizontal symmetry plane from figure Fig. 2.6 and Fig. 2.7, while two symmetry planes means both symmetry planes in Fig. 2.6 and Fig. 2.7.

As seen in Fig. 2.20 to Fig. 2.23, there is a very small deviation between the results of the models which use symmetry planes with respect to the model that does not use the planes. The deviation has maxima close to the frequency of 2 GHz, where both the longitudinal and transverse kick factor have a zero.

The data shows, that symmetry planes may be used to reduce the computational effort of simulations while obtaining correct results. Generally however, care must be taken that no additional components in the model disturb the symmetry. Components that disturb the symmetry are *e.g.*, couplers, vacuum pumps or asymmetrically placed feeding ports. All models subject of this work have the here discussed symmetry and thus, if not stated otherwise, symmetry planes are used.

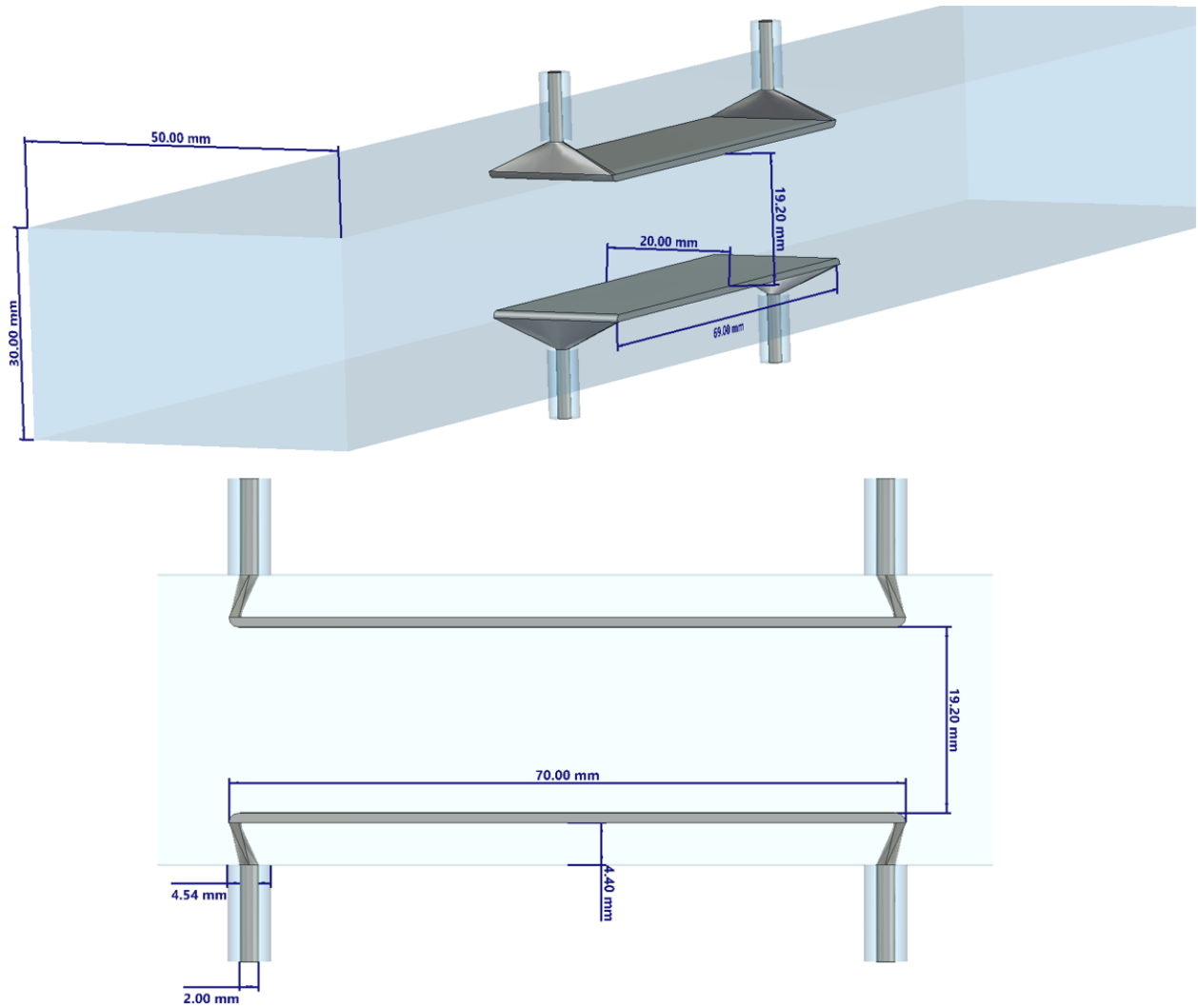


Fig. 2.19: CST model of a stripline kicker designed for a center frequency of 1 GHz.

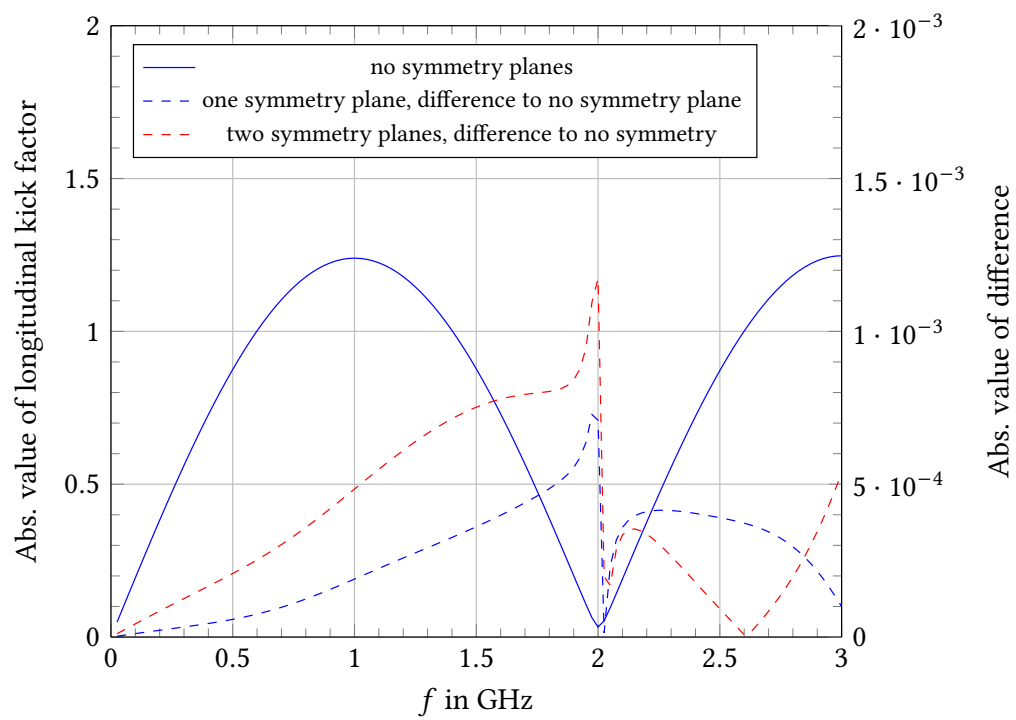


Fig. 2.20: Abs. value of longitudinal kick factor.

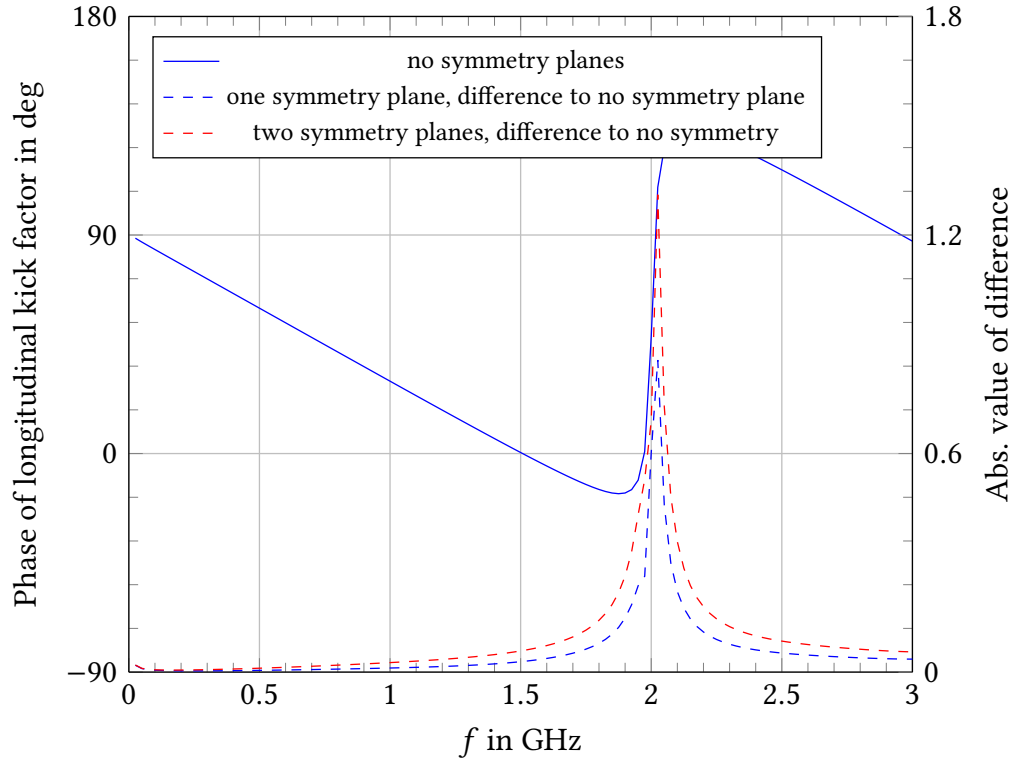


Fig. 2.21: Phase of longitudinal kick factor.

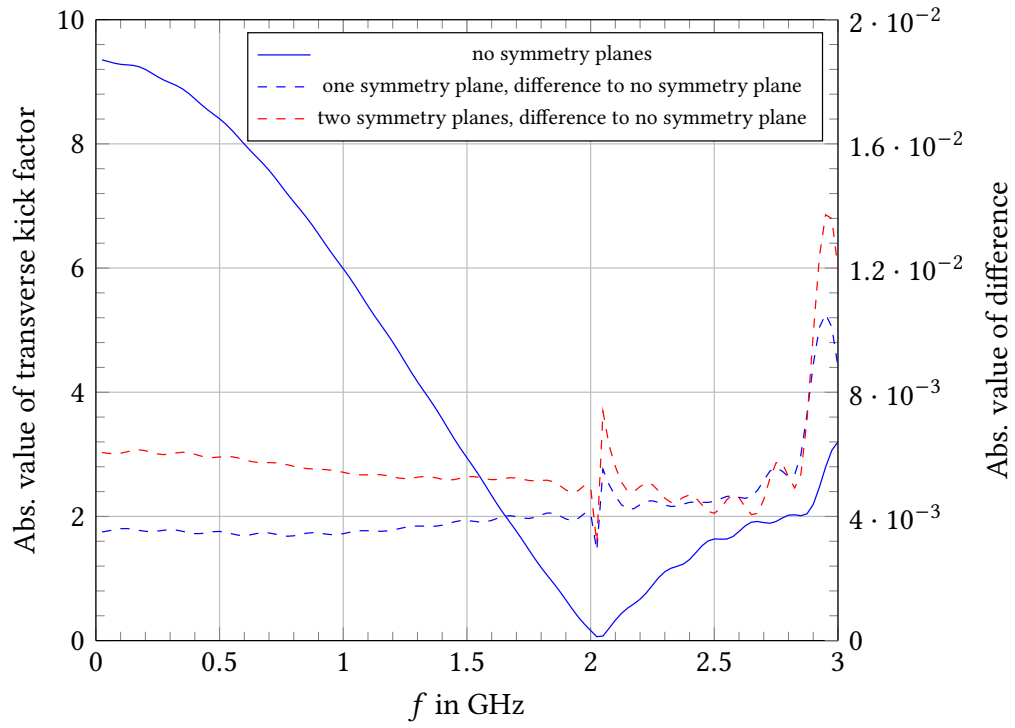


Fig. 2.22: Abs. value of transverse kick factor.

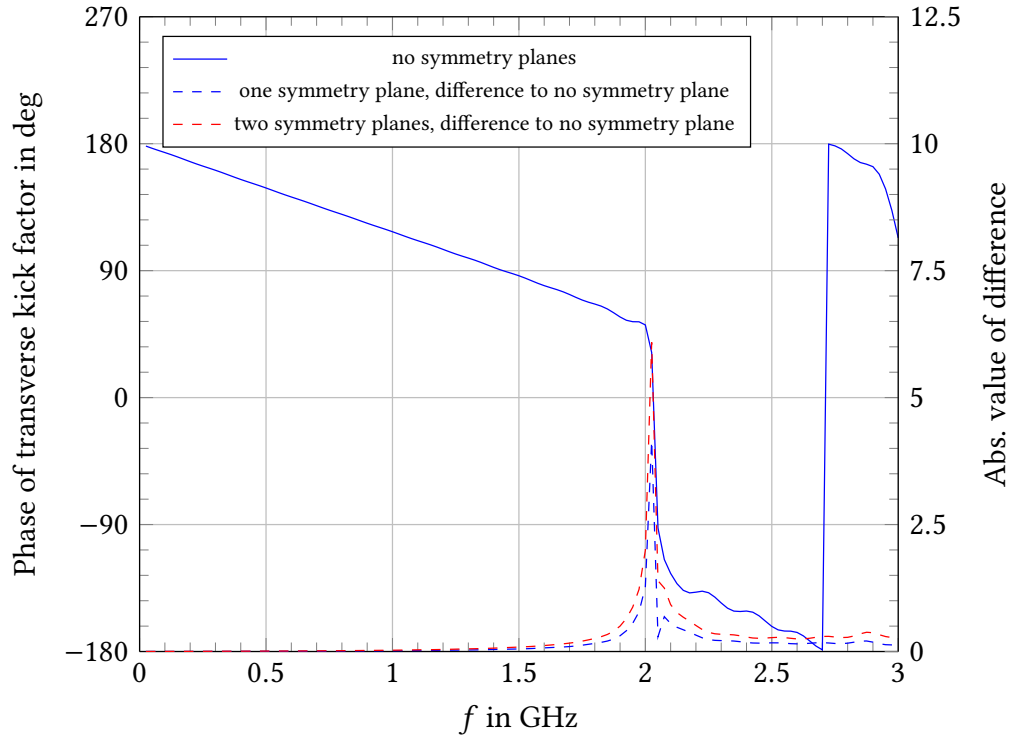


Fig. 2.23: Phase of transverse kick factor.

2.6.2 Numerical analysis using CST

For the computation of wideband kickers, the time-domain and frequency-domain solvers of CST are best suited. By default, the 2024 version of CST includes two tools to compute the longitudinal beam voltage of kicker structures (time-domain solver: VBA Macros»Results»2D 3D Results»Line-Integration of Electric Fields (incl. Transit Time Factor), eigenmode-solver: Template Based Post-Processing»2D and 3D Field Results»3D Eigenmode Results). Another way of computing the beam voltage from CST simulation results is to perform an export of the complex electromagnetic fields and to perform the computation of the beam voltage outside of CST. After the computation of the beam voltage outside of CST, the results of the computation can be imported back into CST, which is useful when performing parameter sweeps or optimization runs.

Both the post-processing macro for the time-domain solver and the post-processing template for the eigenmode solver do not allow computation of the transverse beam voltage. Furthermore, the post-processing macro of the time-domain solver only computes the beam voltage at one given frequency and the computation is very slow. That is why in this work, new CST post-processing templates were developed, that allow the computation of the longitudinal and transverse beam voltages for the time-domain solver [33]. The post-processing templates were developed in the Visual Basic for Applications (VBA) programming language. With the developed post-processing templates the longitudinal and transverse beam voltage (and thus other quantities like kick factor or transfer impedance) can be used for CST parameter sweeps or optimizer runs.

Fig. 2.24 shows the Graphical User Interface (GUI) of the developed post-processing template that computes longitudinal beam voltage (identical for transverse beam voltage). The template computes the complex beam voltage for one or two given values of β and its deviation from linear phase. The nonlinear variation of the phase is the difference of the unwrapped phase and its least-squares fit in a given frequency band. An example of longitudinal beam voltage results of a parameter sweep in CST is shown in Fig. 2.25.

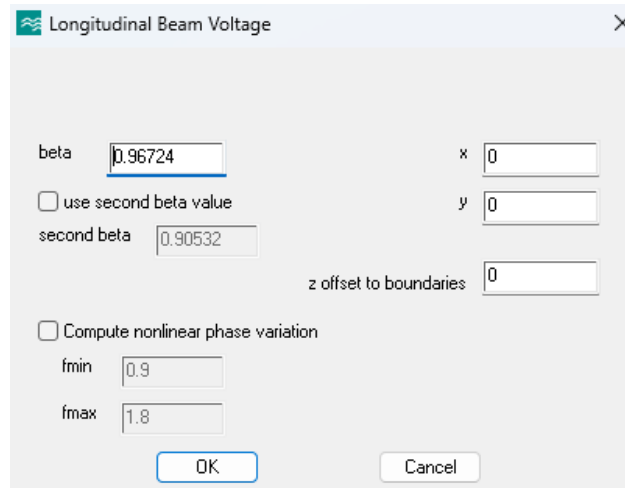


Fig. 2.24: GUI of the developed CST post-processing template for the computation of absolute value and nonlinear phase deviation of the longitudinal beam voltage.

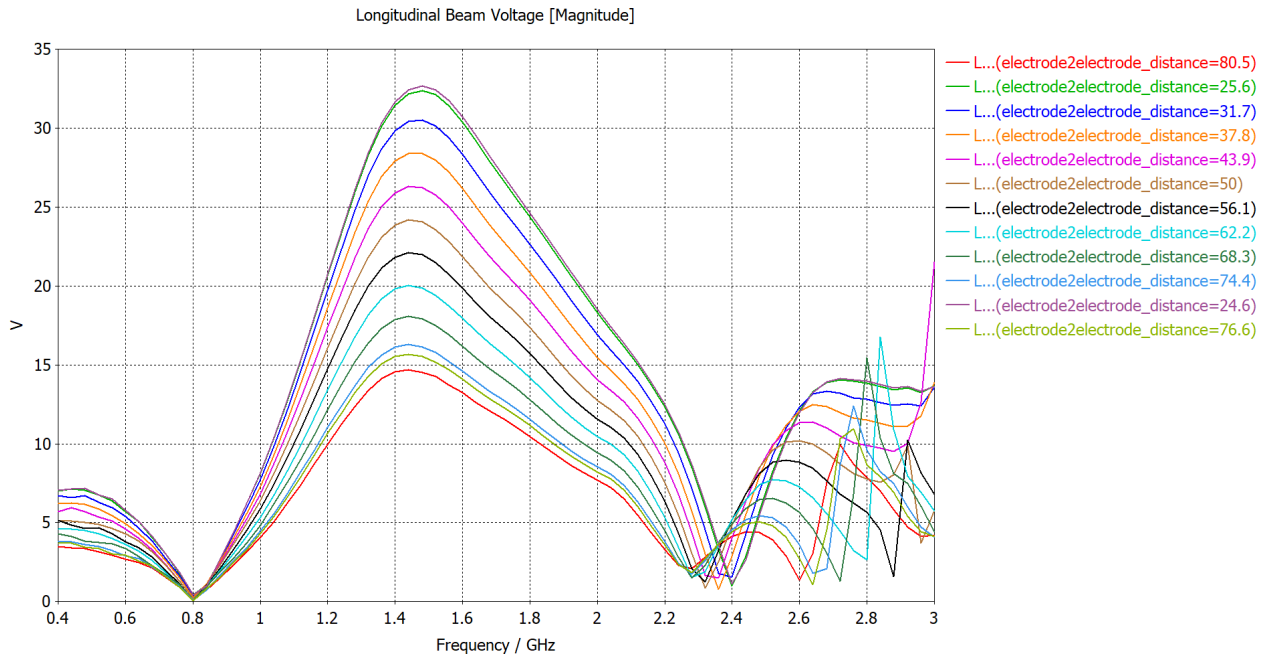


Fig. 2.25: Example for the usage of the developed post-processing template in CST[10]. The abs. value of the longitudinal beam voltage as a result of a parameter sweep of a kicker design is shown.

3 Numerical Simulations of the currently installed Devices

In this chapter, results of CST simulations on the AD stochastic cooling kickers and pickups are shown. The kicker response can be found in Section 3.1 and the pickup response can be found in Section 3.2. Note that both the kicker, and the pickup response depend on the beam velocity, which corresponds to $\beta = 0.96724$ at the first stochastic cooling plateau and $\beta = 0.90532$ at the second stochastic cooling plateau (refer to Fig. 1.2). As mentioned in Chapter 1, there are two kickers and two pickups in the stochastic system of the AD. One pickup-kicker pair is used for stochastic cooling in the horizontal and longitudinal plane and the other pickup-kicker pair is used for stochastic cooling in the vertical and longitudinal plane.

The basic structure of all of the kickers and pickups are arrays of 24 double-loop devices with the same basic dimensions. The arrays have a length of 2.1 m. Sections of the used simulation model can be seen in Figs. 3.1 and 3.2. The ferrites¹ installed in the kicker and pickup structures were taken into account in the simulations. Also, the dielectric discs made out of Vespel®SP-1 material, which hold the inner conductors of the feedline in place, are taken into account. Perfect Electric Conductor (PEC) is used for all metal parts in the simulations.

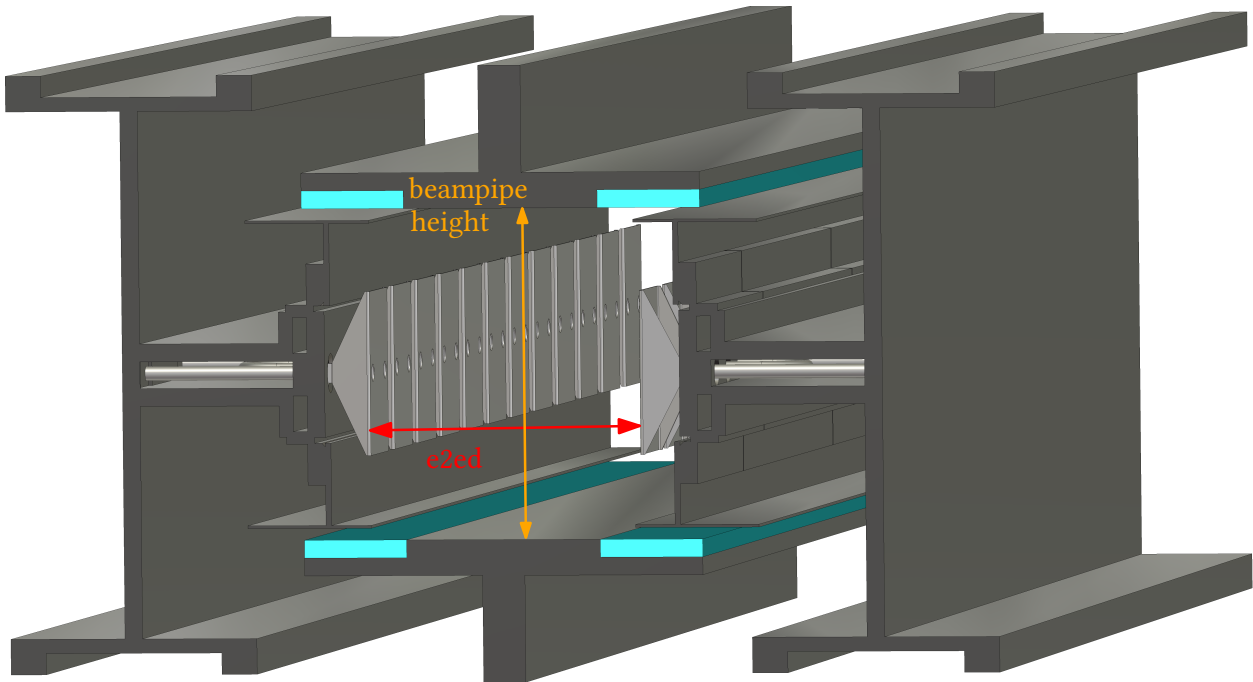


Fig. 3.1: CST model used for the simulations. The ferrites are colored in cyan.

In the pickup and kicker designs, the arrays of double-loops are placed at an angle with respect to the main beam axis. The simulation model is simplified in a sense that the double-loops are not placed at an angle, but the average distance (e2ed) of the double-loops electrodes is used. For all kicker and pickup simulations, beampipe height = 92.9 mm is used (refer to Fig. 3.1).

The original designs of the pickups and kickers allow for movement of the electrodes (variable e2ed in Fig. 3.1). This is used during the cooling, to move the pickup electrodes closer to the beam to increase sensitivity as the beam shrinks. Values of the Electrode to Electrode Distance (e2ed) seen in Fig. 3.1 are listed in Tab. 3.1. For the kicker and pickup that work in the horizontal plane, the minimum and maximum positions of e2ed are marked by KHM in Tab. 3.1, while for the kicker and pickup that work in the vertical plane, the minimum and maximum positions of e2ed are marked by KVM.

¹TT2-111R by Skyworks Solutions, Inc. Formerly known as Transtech Company.

²Documented in the technical drawings PS-C-2176-44-1 and PS-C-0172-44-1.

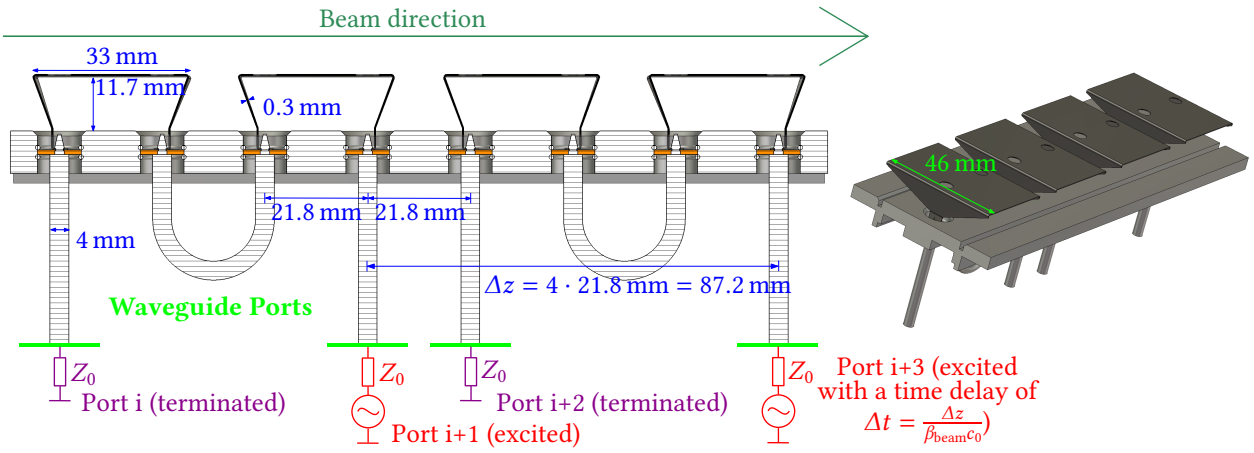


Fig. 3.2: Unit cell of the AD stochastic cooling kicker with feeding scheme and dimensions illustrated. The dimensions are identical for the pickups.

Tab. 3.1: Values of average $e2ed$ in mm.²

$e2ed$ in mm	Notes
24.6	min. position of KVM
25.6	min. position of KHM
31.7	
37.8	
43.9	
50	
56.1	
62.2	
68.3	
74.4	
76.6	max. position of KVM
80.5	max. position of KHM

3.1 Response of the Kickers

The absolute values of longitudinal and transverse kick factor of the currently installed kickers are shown in Fig. 3.3 and Fig. 3.4 for a sweep of the average electrode to electrode distance. Note that since 2014 the movement of the kicker electrodes is no longer used [6]. The electrodes of the kickers have been blocked in the outermost position ever since ($e2ed = 80.5$ mm for the KHM and $e2ed = 76.6$ mm for the KVM).

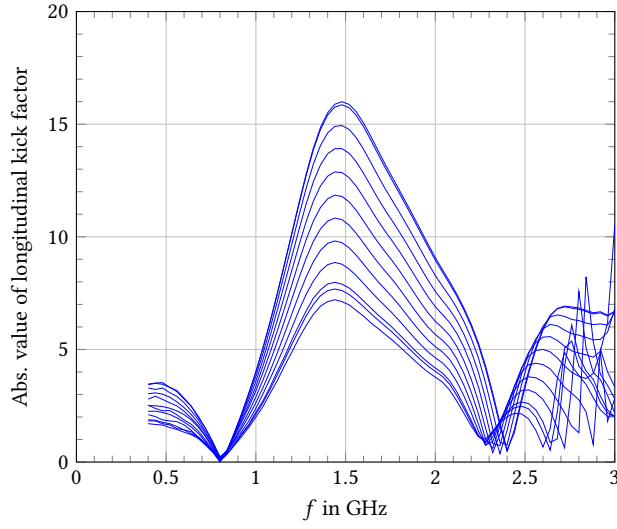


Fig. 3.3: Abs. value of long. kick factor for different values of average $e2ed$ and $\beta = 0.96724$.

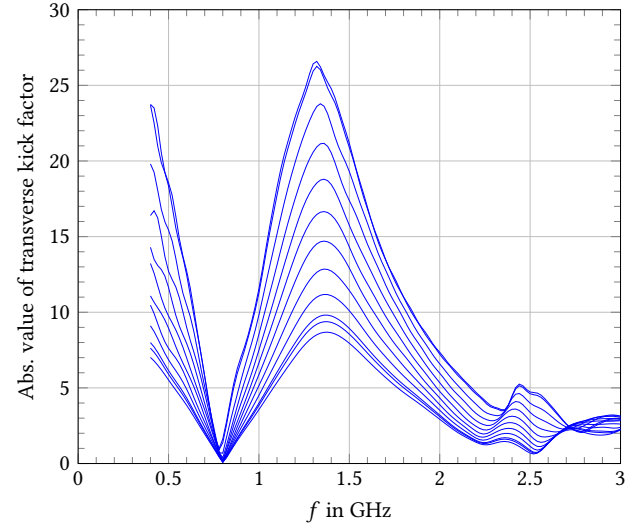


Fig. 3.4: Abs. value of transv. kick factor for different values of average $e2ed$ and $\beta = 0.96724$.

As expected, the zeros at 800 MHz and the position of the maximum do not shift significantly when sweeping the average electrode to electrode distance. It can be observed that the peak value of both the longitudinal and transverse kick factor reduce with the increasing electrode to electrode distance (refer to Fig. 3.5). The spurious behavior above approx. 2.3 GHz is due to waveguide modes propagating inside the beampipe. As seen in Fig. 3.3 and Fig. 3.4, there is a slight asymmetry in the response compared to the behavior expected from the analytical predictions in Section 2.2. This asymmetry can be attributed to the stripline width of 46 mm being a considerable fraction of the wavelength (e.g., $46 \text{ mm}/\lambda = 0.23$ at $f = 1.5$ GHz). A treatment of this effect is given in [34, pp. 75–77] and [35, pp. 640–641]. It appears that the length of the strips (33 mm) does not correspond to a quarter of a wavelength at the frequency where the kicker response has its peak. Like for dipole or patch antennas, the electrical length of the striplines appears to be longer than their physical length. This means that the peak of the response of the stripline will be at a frequency that is lower than expected. This effect is also treated in [34, pp. 75–77] with detailed measurements on stripline pickups with a desired center frequency of 6 GHz.

The deviation of the phase response to a least-squares linear fit in the frequency band from 0.9 GHz to 1.8 GHz is shown in Fig. 3.6 and Fig. 3.7 for the longitudinal and transverse responses. The change of sign at 0.8 GHz can be observed in both the longitudinal and transverse phase response. Furthermore, in the zoomed-in plots of the phase responses in Fig. 3.8 and Fig. 3.9, it can be seen that the nonlinear phase is within $\pm 5^\circ$ in the operational bandwidth for most positions of the average electrode to electrode distance.

Figs. 3.10 and 3.11 show the longitudinal and transverse response of the kickers for maximally opened electrodes. The response is shown for the beam velocities of both cooling plateaus. It can be seen, that the used power splitting scheme (refer to Appendix D) leads to smaller peak value. An improvement in peak value of the kick factor at the second cooling plateau could be obtained by a new design with 24 RF feedthroughs per side of the vacuum tank instead of 12.

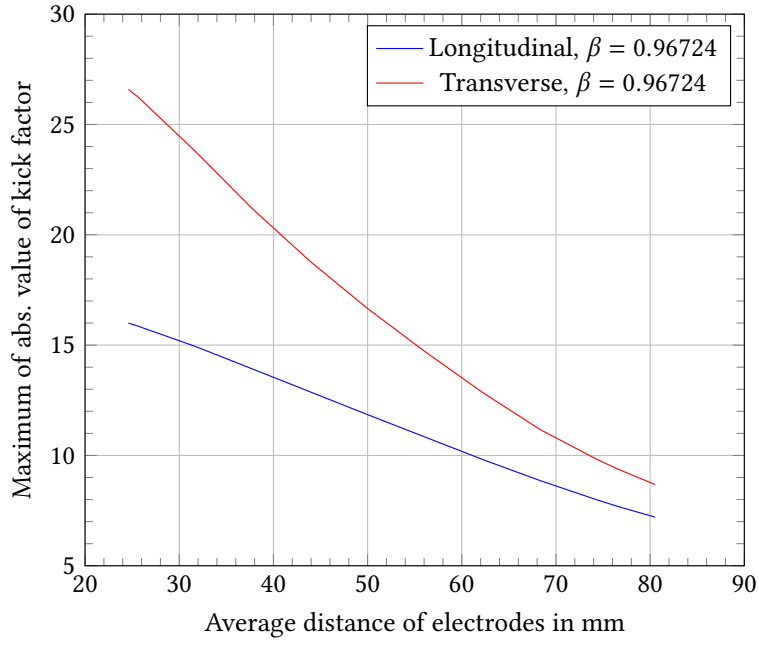


Fig. 3.5: Maximum of abs. value of kick factor vs. average e2ed.

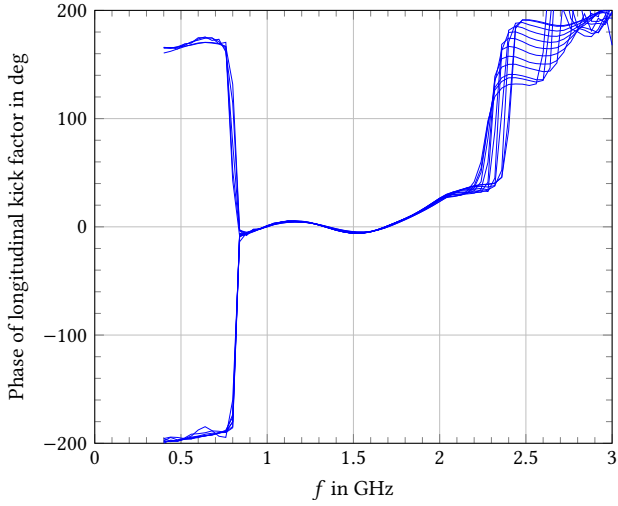


Fig. 3.6: Nonlinear phase of longitudinal kick factor for different values of average e2ed. The nonlinear phase is computed with respect to a linear least-squares fit in the band from 0.9 GHz to 1.8 GHz.

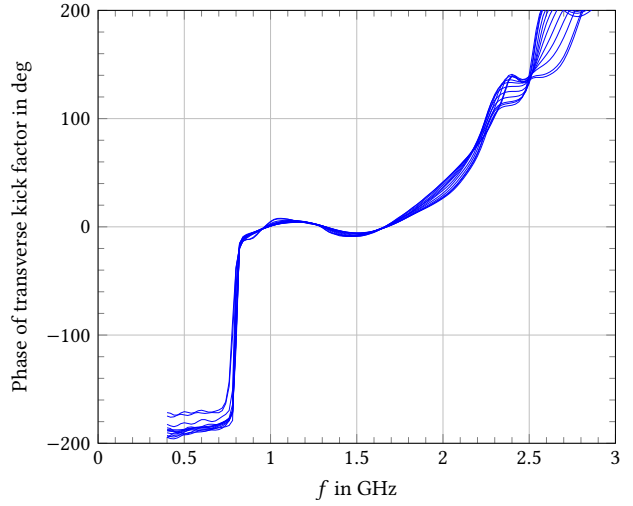


Fig. 3.7: Nonlinear phase of transverse kick factor for different values of average e2ed. The nonlinear phase is computed with respect to a linear least-squares fit in the band from 0.9 GHz to 1.8 GHz.

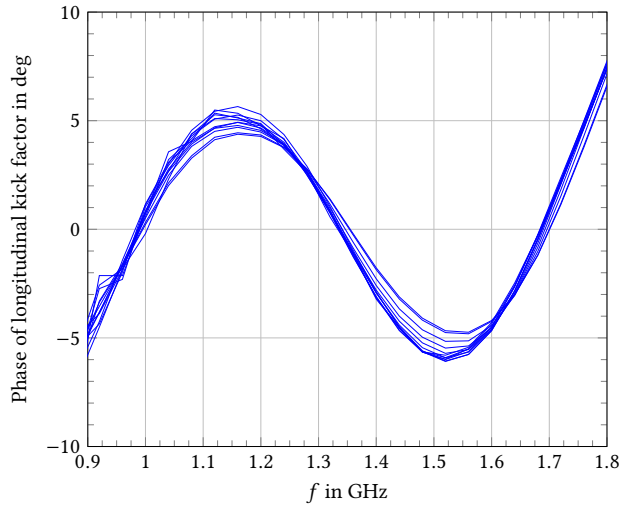


Fig. 3.8: Nonlinear phase of longitudinal kick factor for different values of average $e2ed$. The nonlinear phase is computed with respect to a linear least-squares fit in the band from 0.9 GHz to 1.8 GHz. A zoomed version of Fig. 3.6 is shown.

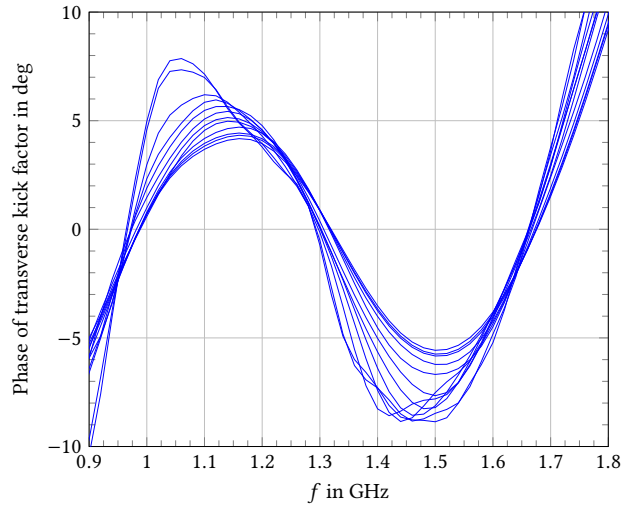


Fig. 3.9: Nonlinear phase of transverse kick factor for different values of average $e2ed$. The nonlinear phase is computed with respect to a linear least-squares fit in the band from 0.9 GHz to 1.8 GHz. A zoomed version of Fig. 3.7 is shown.

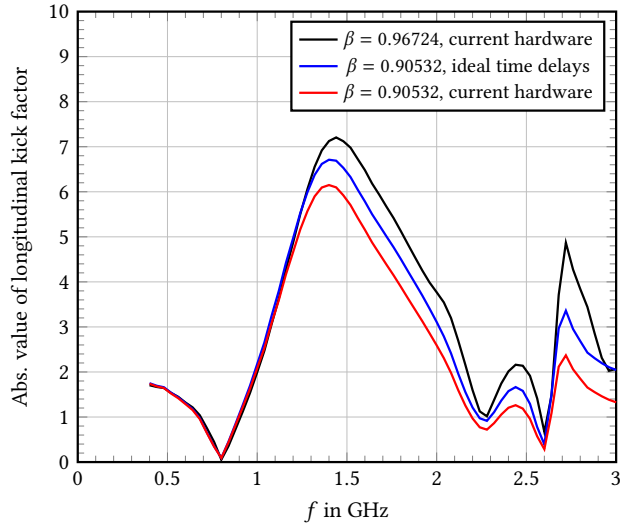


Fig. 3.10: Abs. value of long. kick factor for max opened KHM kicker. Results for the beam velocities at which the stochastic cooling is active are shown.

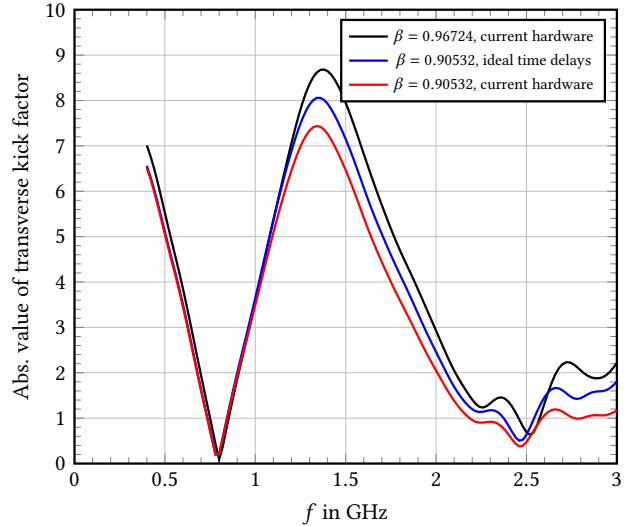


Fig. 3.11: Abs. value of transv. kick factor for max opened KHM kicker. Results for the beam velocities at which the stochastic cooling is active are shown.

3.2 Response of the Pickups

The absolute values of the longitudinal and transverse responses of the pickups are shown in Fig. 3.12 and Fig. 3.13 for $\beta = 0.96724$. The responses are computed from the CST kicker simulation results using the reciprocity relations (2.21) and (2.22) between pickups and kickers.

Like for the kickers, the zeros and maxima of the longitudinal and transverse responses do not shift significantly in frequency when sweeping the average electrode to electrode distance. As seen in Fig. 3.12, the absolute value of the longitudinal transfer impedance varies from 180Ω to 400Ω depending on the positioning of the electrodes. Note that for the longitudinal plane, the total longitudinal transfer impedance varies from $180 \Omega \cdot \sqrt{2}$ to $400 \Omega \cdot \sqrt{2}$ because both pickup tanks deliver longitudinal signals. The transverse transfer impedance varies from $6.5 \text{ k}\Omega/\text{m}$ to $19 \text{ k}\Omega/\text{m}$ (refer to Fig. 3.13). As expected for stripline type devices, the transverse transfer impedance is larger than the longitudinal transfer impedance.

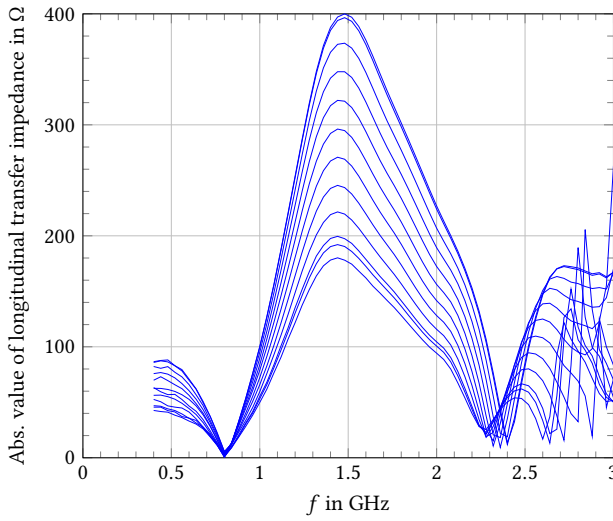


Fig. 3.12: Abs. value of long. transfer impedance for different values of average e2ed.

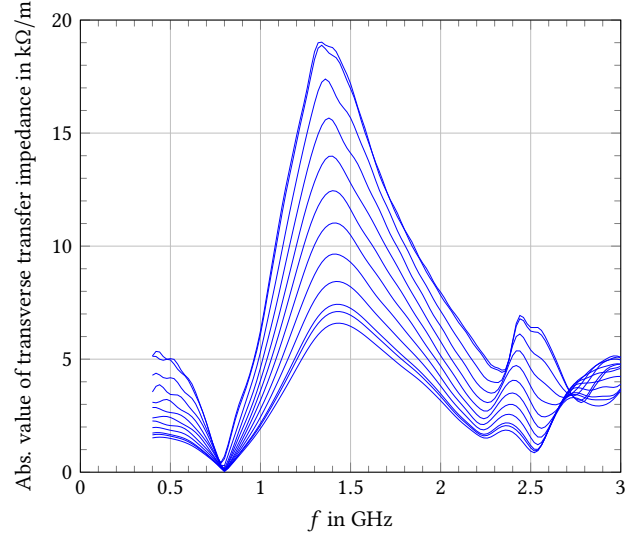


Fig. 3.13: Abs. value of transv. transfer impedance for different values of average e2ed.

From (2.21) and (2.22) follows, that the nonlinear phase of the pickup responses is equal to the nonlinear phase of the kicker. It is thus referred to Figs. 3.6 to 3.9 for plots of the nonlinear phase response of the pickups. A comparison of the pickup behavior at the different stochastic cooling plateaus is shown in Figs. 3.14 and 3.15 for maximally closed electrodes. The pickup tanks only have two signal output ports.³ All signal combination of the 24 double-loops is done inside the vacuum tank for $\beta = 0.96724$. The amplitude of the transfer impedance is thus strongly reduced at the second stochastic cooling plateau ($\beta = 0.90532$). This has been explained in Section 2.4 and the pickup array factor at the second stochastic cooling plateau is the green trace in Fig. 2.17. The maximum of the transfer impedance is plotted against the position of the electrodes in Figs. 3.16 and 3.17.

An improved peak value of the transfer impedance at the second cooling plateau could be obtained, by not doing all signal combination inside the vacuum tank. Like for the kickers⁴, a signal combination scheme could be used, where the signals are not completely combined to only two ports inside the vacuum tank. For example, signal combination to six⁵ instead of two output ports could be done. Switchable delays could then be used outside of the vacuum tank, for combining the signals of these six ports.

³Longitudinal signals are contained in the sum, transverse signals are contained in the difference of the signals from the two ports.

⁴refer to Appendix D for the array factor of the kicker at the second stochastic cooling plateau.

⁵Three per side of the vacuum tank.

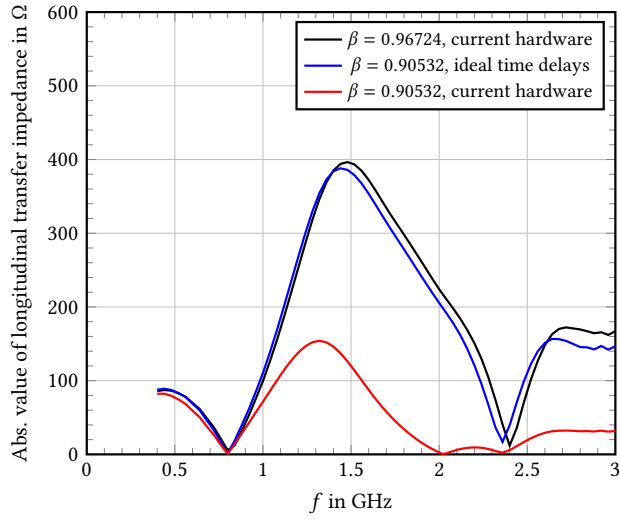


Fig. 3.14: Abs. value of long. transfer impedance for the beam velocities of both cooling plateaus.

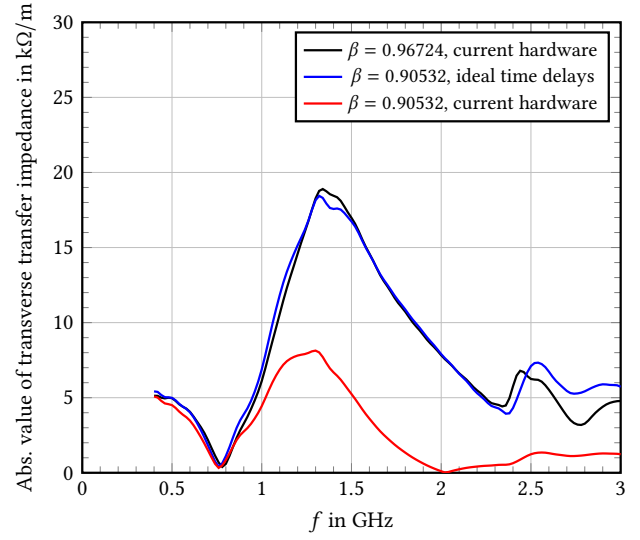


Fig. 3.15: Abs. value of transv. transfer impedance for the beam velocities of both cooling plateaus.

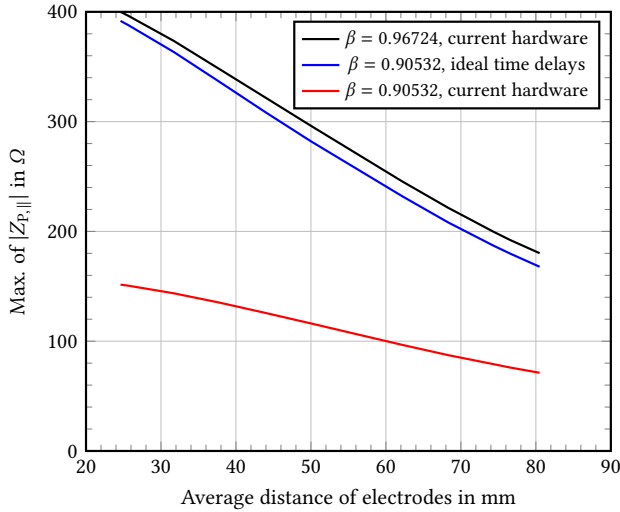


Fig. 3.16: Max. of abs. value of long. transfer impedance for different values of average $e2ed$.

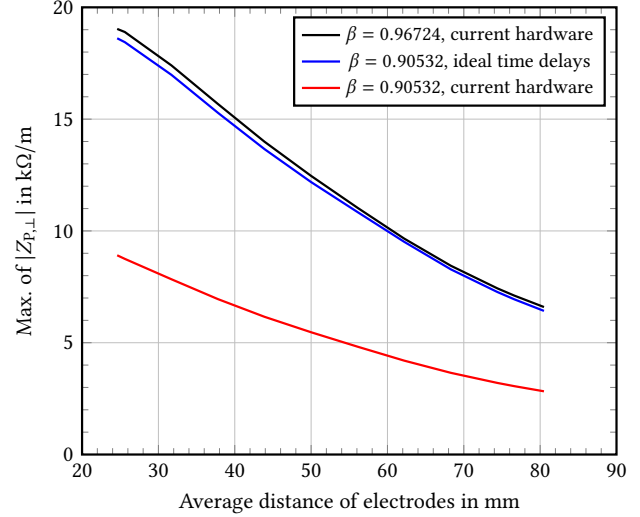


Fig. 3.17: Max of abs. value of transv. transfer impedance for different values of average $e2ed$.

A comparison of the numerically simulated pickup response to measurements is shown in Fig. 3.18. The measurement data is graphically extracted from [11, Fig. 2], where both amplitude- and phase response are shown. This figure shows a "typical frequency response ... for two extreme values of the electrode position corresponding to the beginning and the end of the cooling" [11]. The amplitude scale is in decibel with 5 dB per division without any indication of an absolute level. Due to this, all traces shown in Fig. 3.18 are normalized. In [11], it is not mentioned if the measured response is a longitudinal transfer impedance or a transverse transfer impedance. Thorndahl confirmed, that the measurement is indeed of a transverse pickup response.

It can be seen in Fig. 3.18, that the difference of the maxima of the measured data (black traces) for fully opened- to fully closed array (refer to Tab. 3.1), and the difference of the maxima of the numerically simulated data for fully opened- to fully closed array match reasonably well. This indicates, that the pickups aperture dimensions are similar to those of the KHM, KVM kickers. There is a frequency shift of approx. 150 MHz between the measured and simulated data. The bandwidth of the measured and simulated responses are very similar and the zero at 800 MHz is seen in both the measured and simulated response.

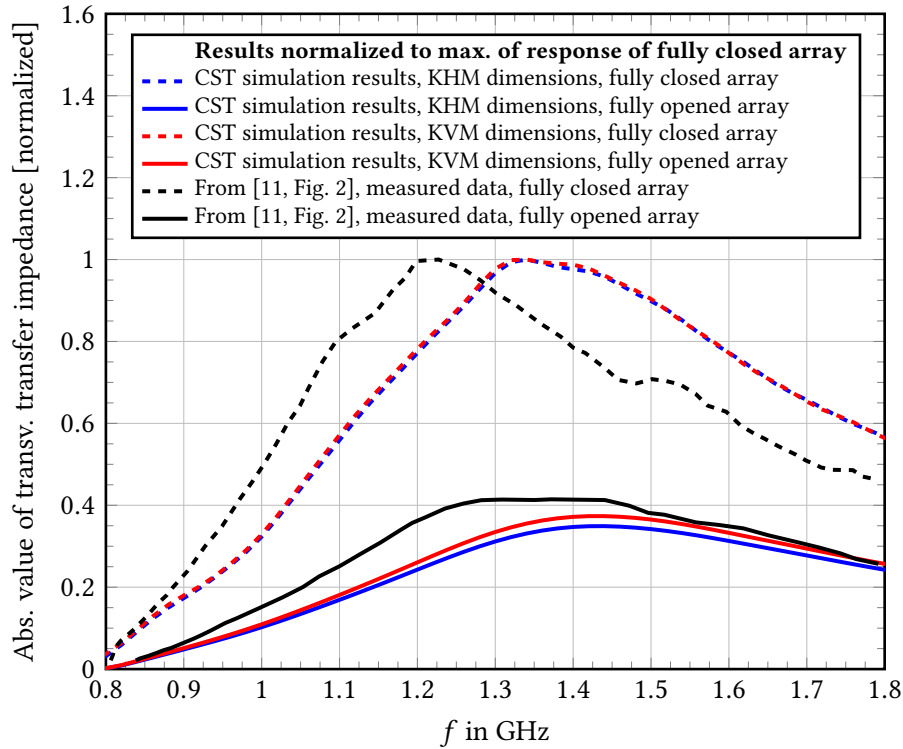


Fig. 3.18: Abs. value of transv. transfer impedance for fully closed- and fully opened double-loop arrays. Comparison of CST simulated response for $\beta = 0.96724$ to measurement data by Autin *et al.*[11]. The measurement data (black traces) is graphically extracted from [11, Fig. 2].

4 Optimizations and alternative Designs

In this chapter, alternative designs which could be used for new stochastic cooling kickers are investigated. The alternative designs either offer higher bandwidth, higher absolute peak value of the kick factor than the currently installed double-loop kickers, or both.

The designs have to fulfill several side-conditions, so that they are integrable in the AD. First of all, the active length¹ of the kicker shall not be larger than 2.1 m, as this is the active length of the currently installed pickups and kickers. Furthermore, the design shall not require power dividers with water-cooled isolation resistors inside the vacuum tank. This eliminates the use of arrays of a large number of resonant slotlines (TSWS), as these types of kick elements are not terminated by loads and thus reflect most of the power fed to them (refer to TSWS in Section 2.3). The signals reflected by the individual slotlines would have to be absorbed by the isolation resistors.

Also, the designs have to be compatible with the currently used power amplifiers and the design should not require major modification of the currently used RF power splitting scheme for feeding power to the kickers. Currently each of the two AD stochastic cooling kickers is powered by 24 power amplifiers with a 3 dB bandwidth from 600 MHz to 1.8 GHz and an output power 1 dB compression point of 100 W [6], [18]. Due to the currently used power amplifiers and power splitting scheme, the kicker designs should not require a number of RF vacuum feedthroughs which is much larger than the currently used number of 24. As a final point, the longitudinal and transverse kick factor should not be substantially lower as those of the currently installed kickers (refer to Chapter 3). Lower kick factor means, that more total RF power is necessary for the same achieved kicks to the antiprotons.

Due to the described side-conditions, the number of individual kick elements contained in the total active length of 2.1 m, should be roughly in the range from 20 to 40. One of the alternative designs is of the well-known slotline topology, whereas two of the alternative designs are of a novel topology which is based on the multi-loop stripline topology. The novel topology is treated in Section 4.1 and the slotline design is treated in Section 4.2.

4.1 Multi-Loops with Broadband Pulse Inverters

The double-loop topology used in the AD stochastic cooling kickers and pickups, uses $\lambda/2$ long delay lines to achieve the necessary 180° phase shift at the center frequency. As seen in the analytical formulas from Section 2.2 and simulation data from Chapter 3, this leads to additional zeros in the amplitude response compared to a single-loop. The increase in peak value has thus to be paid for by a decrease in bandwidth.

From the analytical treatment of double-loops, it can be shown that the additional zeros in the response can be suppressed by using an electrically short pulse inverter instead of a $\lambda/2$ long delay line. Refer to Fig. 4.1 for an illustration of a multi-loop kicker with broadband pulse inverters. A pulse inverter is a circuit which provides a 180° phase shift over a broad bandwidth.

Using broadband pulse inverters with triple-loops makes them competitive to double-loops in terms of bandwidth performance, while having a larger peak value of the kick factor. This approach to multi-loops allows the design of pickups and kickers with higher gain-bandwidth product over the square root of total length than classic double-loops.

In this section, analytical approximations of the kick factor and results of numerical simulations of double- and triple-loops with broadband pulse inverters are shown. The theory and analytical approximations are covered in Section 4.1.1. Results of simulations of simplified inverter double- and triple-loop models are presented in Section 4.1.2. The working principle of a class of broadband pulse inverters is explained in Section 4.1.3. Furthermore, designs of pulse inverters which are suitable for double- and triple-loops within the frequency range relevant to the AD stochastic cooling system are presented (Sections 4.1.4 and 4.1.5). Finally, results of CST simulations of 2.1 m long kickers consisting of arrays of inverter double- and inverter triple-loops are shown (Sections 4.1.6 and 4.1.7).

¹The active length is the length, which is covered by kick elements. A design with 24 individual kick elements with a length of 87.2 mm each has an active length of 2.1 m.

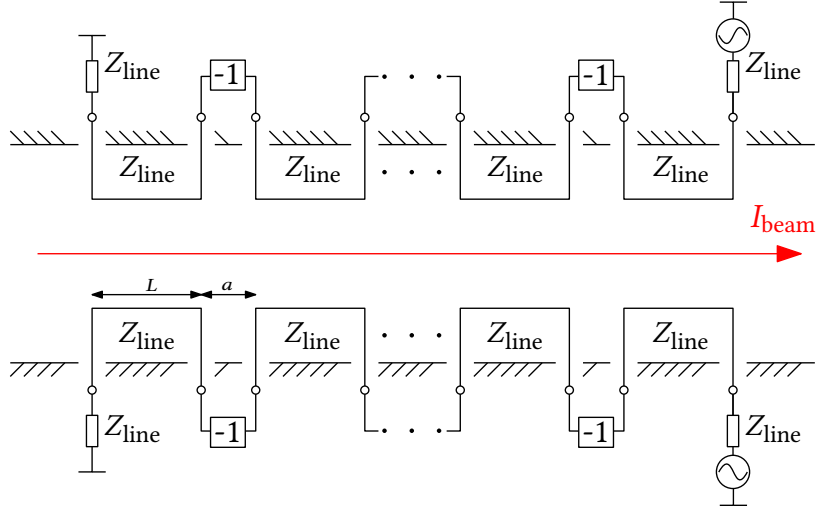


Fig. 4.1: Illustration of a multi-loop kicker with broadband inverters instead of $\lambda/2$ delay-lines. The electrical length of an inverter is b .

4.1.1 Theory

Before stating the intra-loop array factors of double- and triple-loops, it is observed what happens if an ideal inverter circuit is embedded into the $\lambda/2$ long delay lines of a classic double-loop kicker. The comparison of the analytical formulas for the absolute value of the longitudinal kick factor are shown in Fig. 4.2.

Due to the additional 180° phase shift at the center frequency because of the inverter, the individual kicks at the three accelerating gaps cancel and no net kick is applied to a charged particle traversing the kicker. A zero is now introduced at the frequency where the classic double-loop has its maximum. This zero is shifted to higher frequencies if the electrical length of the delay-line is decreased. For the case of zero delay, the zeros of the inverter double-loop align with the zeros of a classic single-loop.

In this case, the inverter double-loop exhibits the same peak amplitude as the classic double-loop configuration, while offering larger bandwidth. The bandwidth is still smaller than the bandwidth of the single-loop. Zero electrical delay is not realizable and for a non-zero electrical delay, there will be a zero between the center frequency and twice the center frequency. From this follows the design goal for the inverters of minimum electrical length. As stated in Section 2.2.4, the complete response of a multi-loop is obtained by multiplying the classic stripline response by the appropriate intra-loop array factor

$$K_{\text{multi-loop}} = K_{\text{single loop}} \cdot \text{AF}_{\text{intra-loop}}. \quad (4.1)$$

(4.1) is used for the traces in Fig. 4.2 and Fig. 4.3. The intra-loop array factors for inverter double- and inverter triple-loops are

$$\text{AF}_{\text{double,inverter}} = 2 \sin(\theta + \zeta) \exp(j(\pi/2 - \theta - \zeta)) \quad (4.2)$$

$$\text{AF}_{\text{triple,inverter}} = 4 \sin(\theta + \zeta + \pi/6) \sin(\theta + \zeta - \pi/6) \exp(j(\pi - 2\theta - 2\zeta)) \quad (4.3)$$

$$\text{with } \theta = \frac{\omega L}{2} \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) \text{ and } \zeta = \frac{\omega}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right). \quad (4.4)$$

The derivations of (4.2) and (4.3) are provided in Appendix C. As seen in Fig. 4.3, an inverter triple-loop exhibits more peak value and bandwidth, than the classic double-loop.

At this point it is important to emphasize, that the zero delay multi-loop results are a best-case upper bound which cannot be achieved in reality because a zero delay connection between the strips is not possible. In the course of this work, it is investigated what is achievable with physically realizable delays. Another important point is the bandwidth of the inverters. In the analytical treatment, the inverters are assumed to be ideal. The bandwidth of the used inverters in the design will thus lower the bandwidth compared to the analytical approximation.

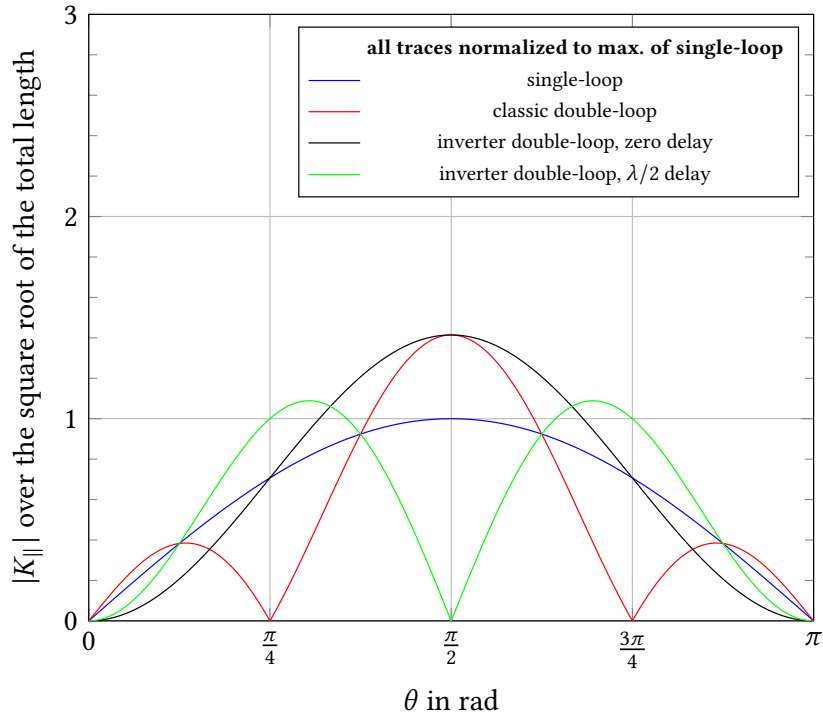


Fig. 4.2: Comparison of analytical formulas for classic single-loops, classic double-loops and inverter double-loops. The longitudinal kick factor over the square root of the total kicker length is plotted against normalized frequency θ . The results are normalized by the maximum value of the response of the single-loop.

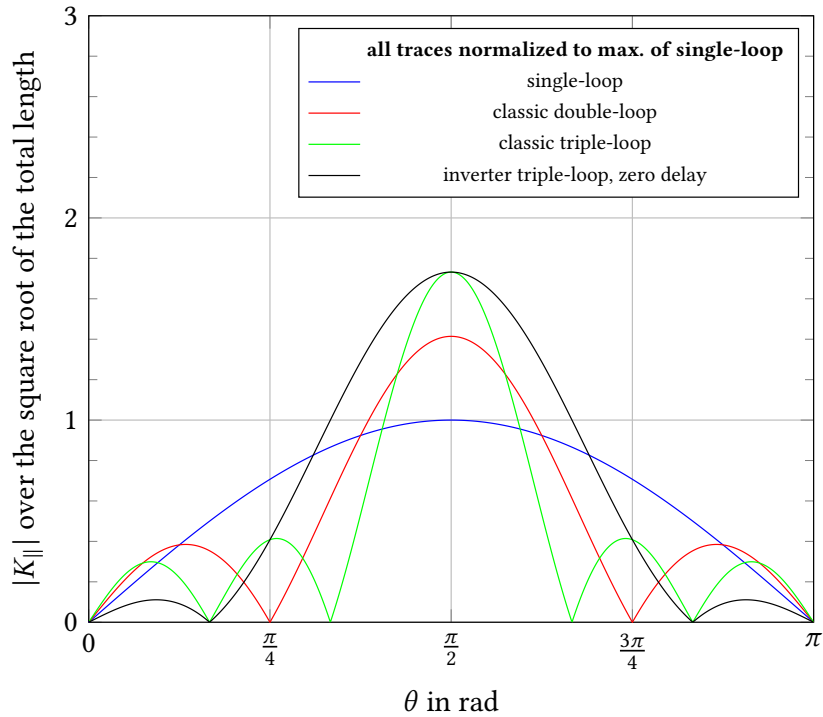


Fig. 4.3: Comparison of analytical formulas for classic single-, double- and triple-loops and the inverter triple-loop. The longitudinal kick factor over the square root of the total kicker length is plotted against normalized frequency θ . The results are normalized by the maximum value of the response of the single-loop.

Before diving into the design of inverter circuits, it is instructive to investigate inverter double-loops and inverter triple-loops for non-zero electrical delays. For this treatment, $v_{\text{beam}} = v_{\text{line}} = c$ is assumed. Assume a classic double-loop and a classic triple-loop which have their maximum absolute value of the kick factor at $\theta = \pi/2$. For such classic multi-loops, the length of the striplines is L and the length of the delay-lines is $2L$. Now suppose, striplines of length αL and inverters of length $\gamma 2L$ are used instead. For the constants α and γ we have $0 \leq \alpha, \gamma \leq 1$. Note, that for the ideal inverter multi-loops in Fig. 4.2 and Fig. 4.3 we have $\alpha = 1$ and $\gamma = 0$.

By using the stripline length αL and electrical inverter length $\gamma 2L$, the results can still be compared to the single-loop or classic multi-loops as functions of the same normalized frequency θ . Following the just described procedure,

$$|K_{\parallel, \text{inverter, double}}| = |\hat{K}_{\parallel, \text{single}}| |\sin(\alpha\theta)| |2 \sin(\theta(\alpha + \gamma))| \quad (4.5)$$

is obtained for the absolute value of the longitudinal kick factor of the inverter double-loop, and

$$|K_{\parallel, \text{inverter, double}}| = |\hat{K}_{\parallel, \text{single}}| |\sin(\alpha\theta)| \left| 4 \sin\left(\theta(\alpha + \gamma) + \frac{\pi}{6}\right) \sin\left(\theta(\alpha + \gamma) - \frac{\pi}{6}\right) \right| \quad (4.6)$$

is obtained for the absolute value of the longitudinal kick factor of the inverter triple-loop. $|\hat{K}_{\parallel, \text{single}}|$ is the maximum of the absolute value of the longitudinal kick factor of a corresponding single-loop. Dividing (4.5) and (4.6) by the square root of the total length gives

$$\frac{|K_{\parallel, \text{inverter, double}}|}{\sqrt{2\alpha L}} = \frac{|\hat{K}_{\parallel, \text{single}}|}{\sqrt{L}} \sqrt{\frac{2}{\alpha}} |\sin(\alpha\theta)| |\sin(\theta(\alpha + \gamma))| \quad (4.7)$$

$$\frac{|K_{\parallel, \text{inverter, triple}}|}{\sqrt{3\alpha L}} = \frac{|\hat{K}_{\parallel, \text{single}}|}{\sqrt{L}} \sqrt{\frac{4}{3\alpha}} |\sin(\alpha\theta)| \left| \sin\left(\theta(\alpha + \gamma) + \frac{\pi}{6}\right) \sin\left(\theta(\alpha + \gamma) - \frac{\pi}{6}\right) \right|. \quad (4.8)$$

Plots of (4.7) for different values of α and γ are shown in Fig. 4.4 and plots of (4.8) for different values of α and γ are shown in Fig. 4.5. It can be seen, that for constant length of the striplines ($\alpha = 1$) and increasing electrical delay of the inverter (γ from 0 to 0.6), the position of the maxima are shifted to lower frequencies. This shift to lower frequencies can be counter-acted by decreasing the length of the striplines. This can be seen on the blue traces in Fig. 4.4 and Fig. 4.5.

Comparing the blue and black traces in Fig. 4.4 and Fig. 4.5, the following can be concluded: Inverter multi-loops with realistic electrical delays² of the pulse inverters can be made to have the same center frequency and approximately the same maxima of kick factor over the square root of the total length than the ideal inverter multi-loops by reducing the length of the strips. The inverter multi-loops with realistic delays then have larger gain-bandwidth product over square root of total length than the corresponding classic multi-loops.

The following example of the double-loop illustrates this: An inverter double-loop with strips of 60% strip length compared to a classic double-loop and with inverters of 60% electrical length compared the delay-lines of the classic double-loops has the same peak value but larger bandwidth than the classic double-loop. This can be seen in Fig. 4.6 and an analogous example for the inverter triple-loop is shown in Fig. 4.7.

² $\gamma = 0.6$ can be considered a realistic time-delay. The electrical delay of the inverter would have to be 60% of that of a classic double-loop.

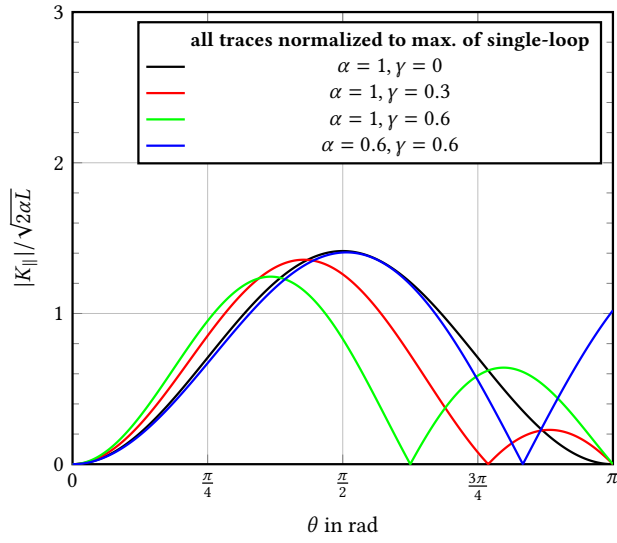


Fig. 4.4: Inverter double-loop response for different values of α & γ .

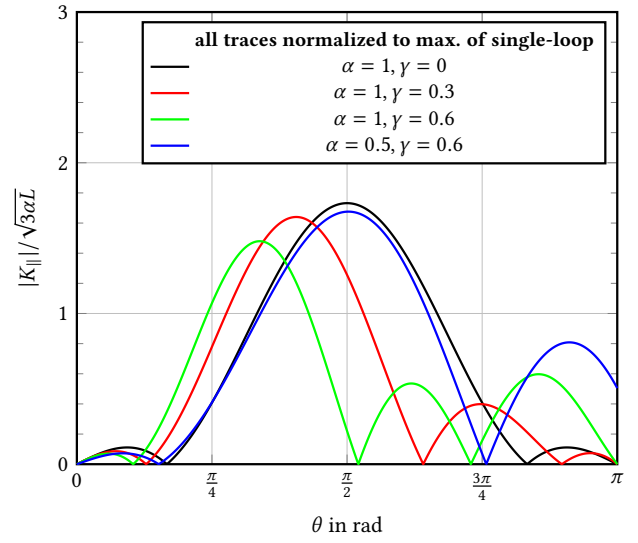


Fig. 4.5: Inverter triple-loop response for different values of α & γ .

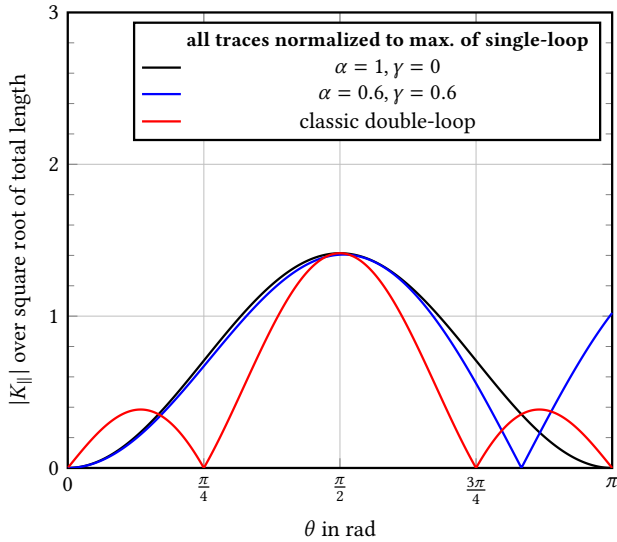


Fig. 4.6: Inverter double-loop response for different values of α & γ . Comparison to classic double-loop response.

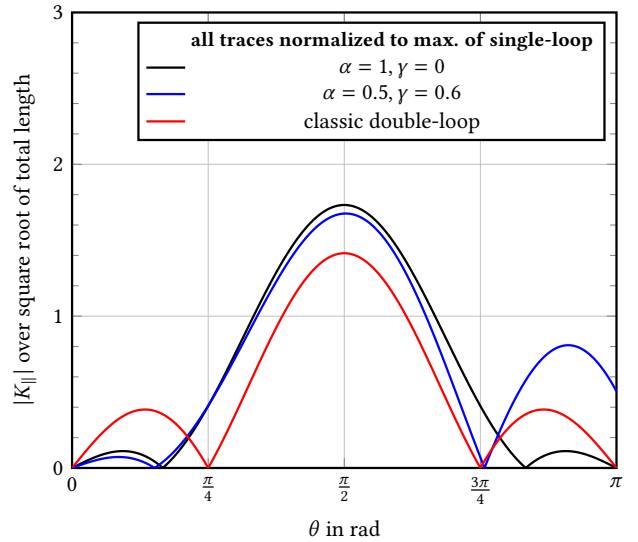


Fig. 4.7: Inverter triple-loop response for different values of α & γ . Comparison to classic double-loop response.

4.1.2 CST simulated Performance of Inverter Double-Loops and Inverter Triple-Loops

In this subsection, results of CST simulations of inverter double- and inverter triple-loops are shown. In this first step of numerical simulations, instead of full 3D Electromagnetic (EM)-simulations of multi-loop kicker arrays with broadband inverters, simplified models were used. These simplified 3D-models consist of classic stripline arrays (refer to Figs. 4.8 and 4.9), which have the same dimensions as the current AD stochastic cooling kicker double-loops. Each stripline has two coaxial waveguide ports. In CST transient co-simulations, double- and triple-loops have been created by connecting adjacent striplines via ideal delay-line circuit elements and ideal inverter circuit elements.³

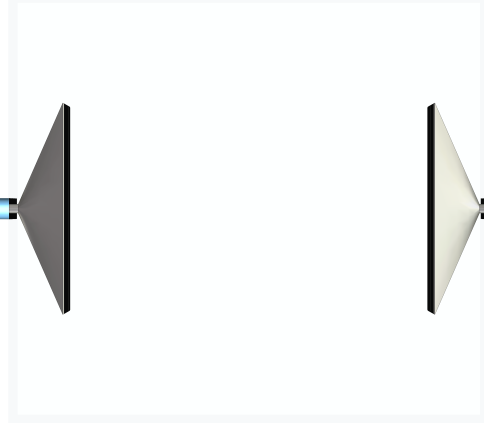


Fig. 4.8: Cross-section view of CST model, used for the co-simulations. An xy -plane view is shown.

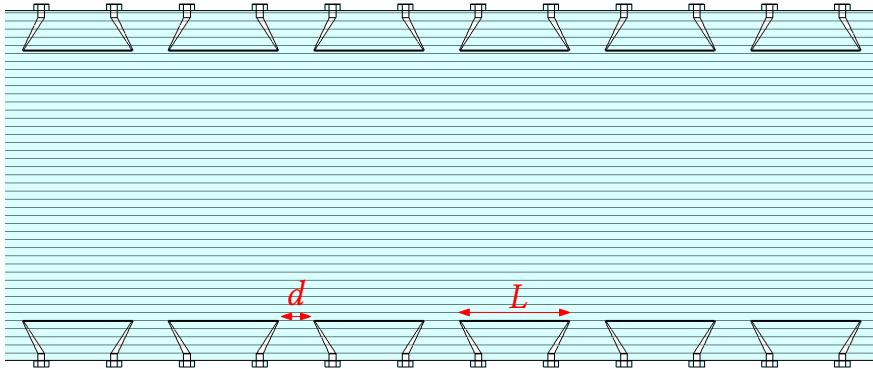


Fig. 4.9: Cross-section side-view of CST model, used for the co-simulations.

With this setup, the CST time-domain solver solves Maxwell's equations only on the domain of the beampipe with the stripline elements instead of also having to compute the EM-fields inside the inverter structures. The interaction of the waveguide-ports with the inverters is simulated with the transient circuit solver of CST. This allows for fast simulations.

The typical beam-aperture dimensions of 80.5 mm electrode to electrode distance and 92.9 mm beam-pipe height were used. In this investigation, ferrites were not used. In addition to inverter double- and inverter triple-loops, a classic double-loop kicker was simulated. For this classic double-loop kicker, ideal circuit model transmission lines have been used for the co-simulation. All simulated models share the same geometry of the striplines and the same beam-aperture. The only parameters that change are the phase shift of the inverters, the electrical delay-line length, the strip length L and the strip-to-strip distance d . This way, a faithful comparison of the simulation results of the models is possible.

³In CST transient co-simulations, circuit models of several components (*i.e.*, lumped elements, transmission lines, junctions, splitters, phase shifters) can be used, which can be connected to TEM-waveguide ports of 3D models. The ideal phase-shifter and transmission line of defined electrical length were used to model the broadband 180° phase shifters with non-zero electrical delay.

The simulation results for longitudinal- and transverse kick factor of the classic double-loop, inverter double-loop and inverter triple-loop are shown in Fig. 4.10 and Fig. 4.11. The first zero in the response is suppressed by the inverter designs. The gain in bandwidth for the inverter double-loop is paid for by a small reduction of peak value. The inverter triple-loop offers more bandwidth and peak value than the classic double-loop in both the longitudinal and the transverse response. As seen in Fig. 4.11, the large DC peak in the transverse response is suppressed by the inverter designs. A comparison of gain-bandwidth-product over square root of length $\text{GBWP}/\sqrt{L}$ is shown in Tab. 4.1. The inverter double-loop has a gain of 15% in the longitudinal $\text{GBWP}/\sqrt{L}$ and 21% in the transverse $\text{GBWP}/\sqrt{L}$ compared to the classic double-loop. The inverter triple-loop has a gain of 22% in the longitudinal $\text{GBWP}/\sqrt{L}$ and 28% in the transverse $\text{GBWP}/\sqrt{L}$ compared to the classic double-loop.

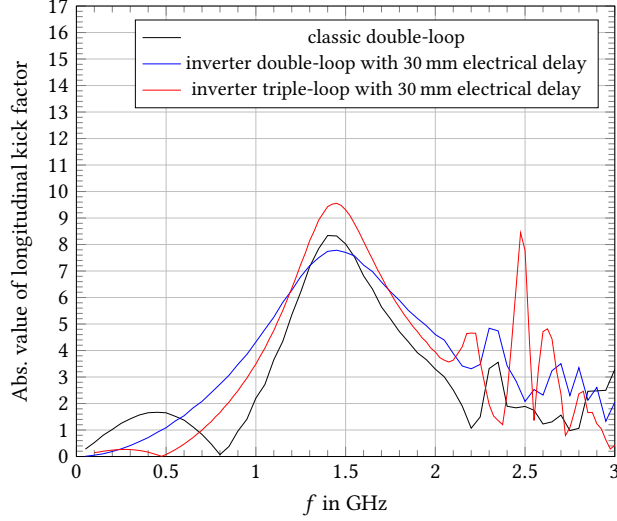


Fig. 4.10: Abs. value of longitudinal kick factor for $\beta = 0.96724$ of a 2.1 m long kicker.

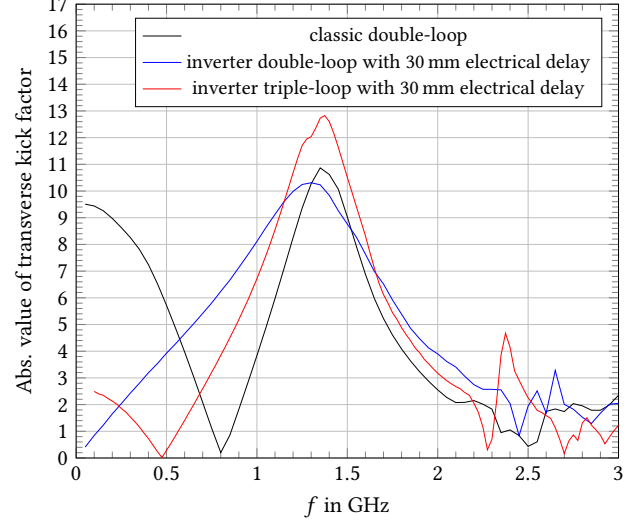


Fig. 4.11: Abs. value of transverse kick factor for $\beta = 0.96724$ of a 2.1 m long kicker.

Tab. 4.1: Gain-bandwidth-product over square root of length in $\text{GHz}/\sqrt{\text{m}}$. The raw data is from Fig. 4.10 and Fig. 4.11. Integration is done from 0.9 GHz to 1.8 GHz.

	longitudinal	transverse
classic double-loop	3.44	4.37
inverter double-loop	3.94	5.29
inverter triple-loop	4.18	5.58

4.1.3 Broadband Pulse Inverters

Fig. 4.12 illustrates the requirements for the broadband pulse inverters for use in multi-loops. Apart from the trivial requirements of low reflection coefficients and low insertion loss in the operational bandwidth, the electrical length should be as small as possible. Furthermore, as we are interested in the design of kickers, the devices have to be compatible with the power levels needed to achieve the required beam voltages. Typically, there is a conflict between the latter two requirements. In the design of broadband pulse inverters using coaxial transmission lines, it has been found that small electrical length is in conflict with achieving bandwidth, insertion loss and return loss requirements.

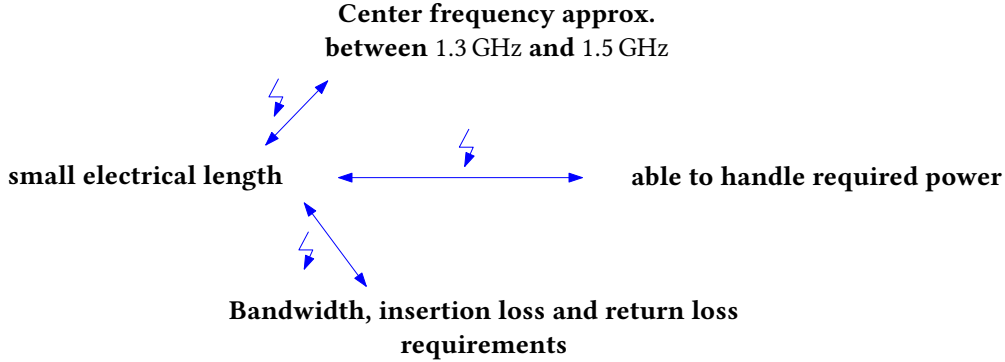


Fig. 4.12: Illustration of the requirements of a broadband pulse inverter for multi-loop kickers.

For the design of pickups, the requirement of power handling capability is eliminated as we are dealing with very low intensity beam and thus very small power levels. This also allows the use of smaller structures which makes achieving small electrical length easier. One requirement which remains very important is insertion loss, as any dissipative loss introduced by the inverter degrades the Signal-to-Noise Ratio (SNR) of the total pickup. This makes the use of lossy materials to increase the bandwidth mostly unviable. This is due to the fact that a gain of sensitivity by employing a triple- or quadruple-loop design would be followed by increased noise levels, which may lead to no improvement or in fact a degradation of the SNR of the total pickup.

At microwave frequencies, there are a multitude of possible realizations of broadband pulse inverters. Coupled transmission line circuits [36]–[38], microstrip to coplanar waveguide to microstrip transitions [39] and microstrip to slotline to microstrip transitions [40, Chapter 6],[41], [42] are some of the possible realizations. All of these designs are realized on a dielectric substrate. This makes achieving very small electric lengths such as the required 30 mm from Fig. 4.10 and Fig. 4.11 very hard, keeping in mind that the inverter circuit needs to mechanically connect striplines in the kicker.

A particularly simple realization of an RF pulse inverter is a coaxial transmission line, where at one point inner and outer conductors are flipped. This topology is very popular in RF baluns for feeding antennas. Another application of coaxial-line inverters was high-speed cathode ray tube oscillography [43]. In this topology, the lower cut-off frequency of the operational bandwidth can be extended down to frequencies in the single-digit MHz range by using ferrites [44].

A simple CST-model is shown in Fig. 4.13, where the boundary is grounded. The broadband 180° phase shift is achieved by the electric field lines changing direction due to flipping inner- and outer conductor of the 50 Ω coaxial line. This can be seen in the plot of the electric field in Fig. 4.14.

In the plots of the magnitude of reflection- and transmission coefficient (Fig. 4.15 and Fig. 4.16), it can be seen that the response is good if the bounding box is large. To achieve low delay, the inverters will be placed close to other conductive parts of the kicker. As seen in the plots, a configuration with low time-delay (a configuration with small dimensions of the bounding box) yields bad transmission and matching behavior in the frequency band relevant for the AD stochastic cooling system.

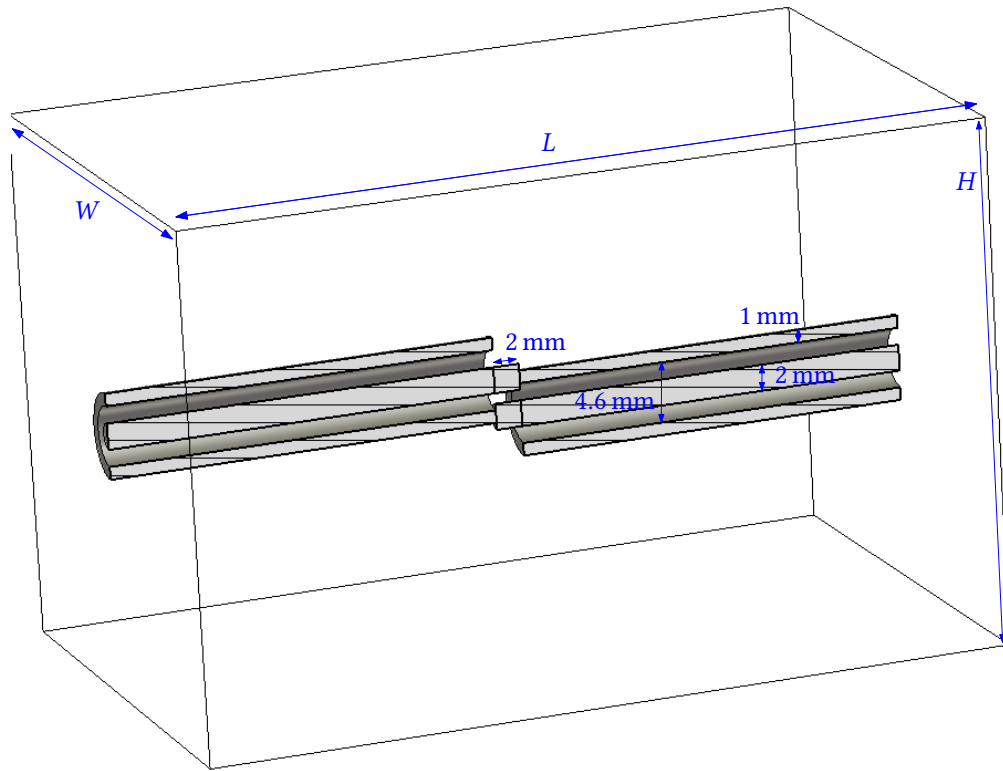


Fig. 4.13: 3D view of the cross-section of a simple coaxial-line inverter. The boundary box is grounded. The grey parts are PEC and the background material is vacuum.

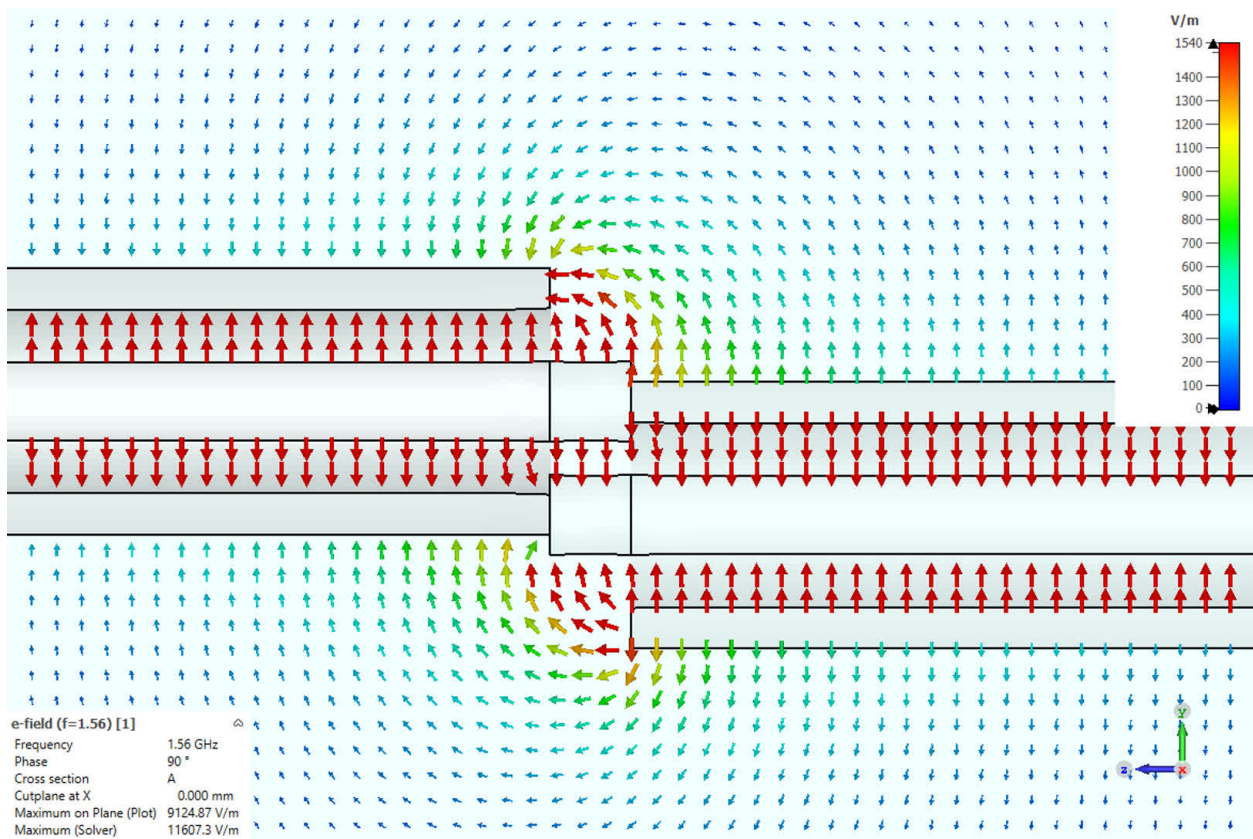


Fig. 4.14: E -Field at 1.56 GHz in cross-section of the coaxial-line inverter for $W = H = 40$ mm, $L = 62$ mm.

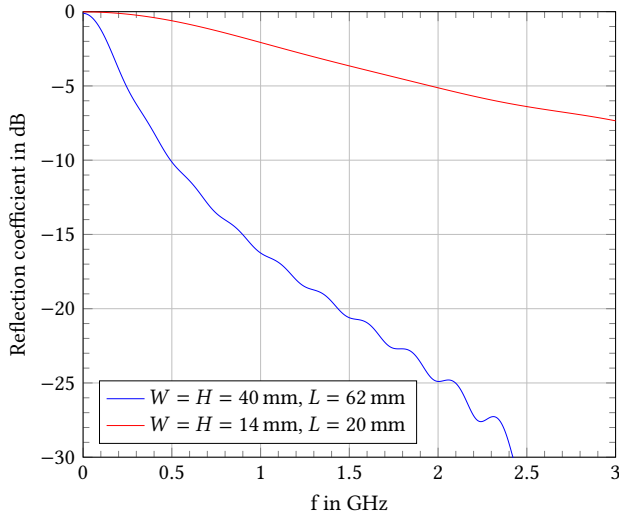


Fig. 4.15: Reflection coefficient of the coaxial pulse inverter.

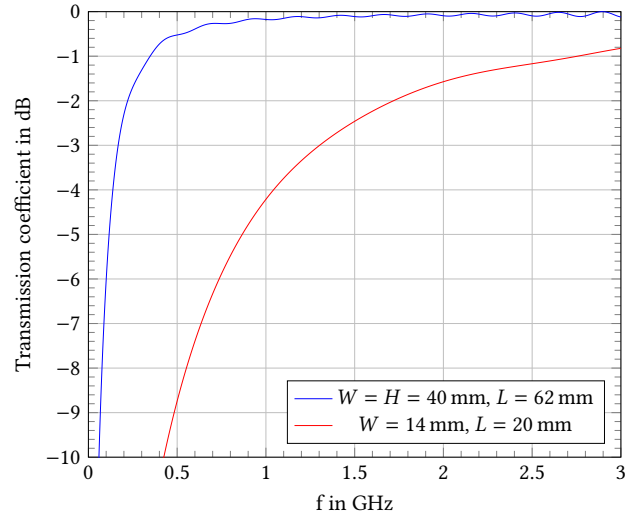


Fig. 4.16: Transmission coefficient of the coaxial pulse inverter.

Article [43] by Rochelle contains a comprehensive treatment of coaxial-line inverters with conductive parts (shielding) in vicinity of the inverter. The most relevant points of this treatment are summarized in the following. The model includes shielding of the area where inner and outer conductor are flipped (refer to Fig. 4.17). The shielding itself forms a coaxial transmission line with the outer conductor of line 2.

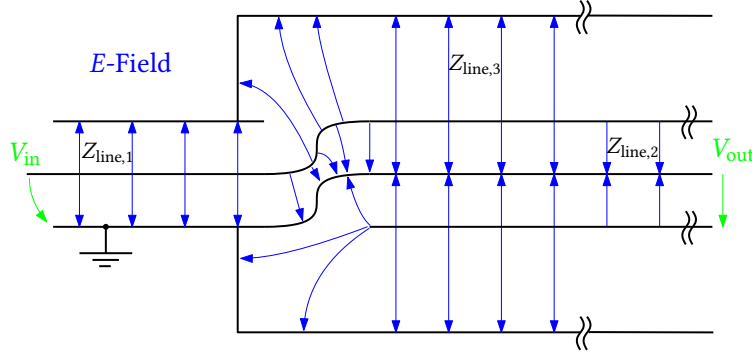


Fig. 4.17: Illustration of a coaxial pulse inverter. Lines 2 and 3 are terminated by their line impedance. Based on [43, Fig. 1, Fig. 2].

Assuming line 1 is fed by a matched source, and lines 2 and 3 are terminated by their line impedance, the system can be modelled by a transmission line of impedance $Z_{\text{line},1}$, terminated by the parallel circuit of $Z_{\text{line},2}$ and $Z_{\text{line},3}$. The output and input voltage pulses are thus related by⁴

$$V_{\text{out}}(t) = -V_{\text{in}}(t) \cdot \left(1 + \frac{\frac{Z_{\text{line},2}Z_{\text{line},3}}{Z_{\text{line},2}+Z_{\text{line},3}} - Z_{\text{line},1}}{\frac{Z_{\text{line},2}Z_{\text{line},3}}{Z_{\text{line},2}+Z_{\text{line},3}} + Z_{\text{line},1}} \right) = -2V_{\text{in}}(t) \cdot \frac{Z_{\text{line},3}Z_{\text{line},2}}{Z_{\text{line},3}Z_{\text{line},2} + Z_{\text{line},3}Z_{\text{line},1} + Z_{\text{line},2}Z_{\text{line},1}}. \quad (4.9)$$

Assuming $Z_{\text{line},2} = Z_{\text{line},1}$, we get

$$V_{\text{out}}(t) = -V_{\text{in}}(t) \cdot \frac{2 \frac{Z_{\text{line},3}}{Z_{\text{line},1}}}{2 \frac{Z_{\text{line},3}}{Z_{\text{line},1}} + 1}. \quad (4.10)$$

From (4.10), which relates the desired output pulse to the input pulse follows, that in order to keep the amplitude of the desired output pulse high, $Z_{\text{line},3}$ should be as high as possible. The formula of the line

⁴The time-delay from input to output is neglected. (4.9) is plotted in [43, Fig. 4]

impedance of a coaxial transmission line is

$$Z_{\text{line,coaxial}} = \frac{1}{2\pi} \cdot \sqrt{\frac{\mu}{\epsilon}} \cdot \ln\left(\frac{r_{\text{outer}}}{r_{\text{inner}}}\right) \quad (4.11)$$

and it is thus beneficial to keep the distance between shield and line 2 as large as possible. In addition to this rather trivial result, the impedance of line 3 can be also boosted by placing a material of high μ/ϵ -ratio in the shield.

Let line 3 now be terminated by a short circuit at a distance L from the transition (as seen in Fig. 4.18). If $v_{\text{phase,line3}}$ is the phase velocity of line 3, then additionally to the wanted inverted voltage pulse at the output, there will be a second pulse at a time $2L/v_{\text{phase,line3}}$ after the desired inverted pulse. The passband of the device in frequency domain is around the frequency where $L = \lambda/4$, as the short circuit at the end of the shielding transforms to an open at the point of the cross-connection of the conductors. The bandwidth of the device can be additionally boosted by embedding a lossy material in the shield as this damps the amplitude of the undesired echo pulse.

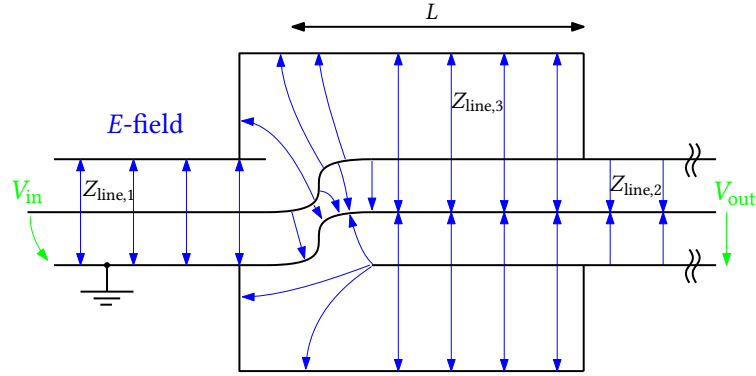


Fig. 4.18: Illustration of a coaxial pulse inverter. Lines 3 is terminated by a short. Line 2 is terminated by its line impedance. Based on [43, Fig. 1, Fig. 2].

A folded version of the coaxial-line inverter is more promising (refer to Fig. 4.19) for integration in a multi-loop kicker design, as in this configuration it is easier to connect the striplines to the coaxial-lines. This configuration is a coaxial-line to two-wire line to coaxial-line transition. The two-wire line coupling the coaxial-lines has the impedance $Z_{\text{line,3}}$. Here, the parasitic transmission line is the two-wire transmission line of impedance $Z_{\text{line,4}}$, formed by the outer conductors of the coaxial-lines. For good transmission behavior, it is thus beneficial, if the coaxial-lines are placed far apart, to have a high μ/ϵ -ratio between the two coaxial-lines and/or to have a lossy material in between the coaxial-lines. The latter may improve bandwidth and matching but introduces insertion loss, which is definitely not wanted in our application.

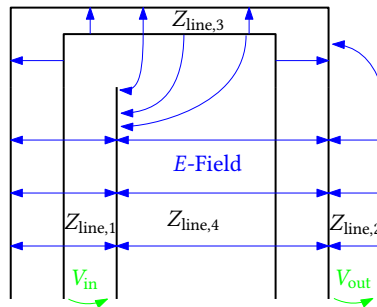


Fig. 4.19: Illustration of a folded coaxial-line inverter. Based on [43, Fig. 1].

4.1.4 A Pulse Inverter Design for a Double-Loop Kicker

A pulse inverter design suitable for a double-loop kicker is presented. The details of its mechanical design are shown and the electromagnetic transmission behavior investigated. Fig. 4.20 shows a blown-up 3D-view of the inverter. Side-views of the model are shown in Figs. 4.21 to 4.24. A coaxial-line to parallel-plate waveguide to coaxial-line transition is chosen as the basic topology. As seen in the blown-up view of the inverter structure, the parallel-plate waveguide is formed by two metal sheets, which may be made from the same material as the striplines. The base of the structure may be CNC-machined out of aluminum. For the simulations, PEC has been used as material.

A plot of the electric fields in a cross-section of the inverter is shown in Fig. 4.25 at the frequency of 1.4 GHz. Like in the introductory example, the flipped direction of the electric field lines on the two coaxial-lines can be observed, which leads to the broadband 180° phase shift. Optimization runs using the CST frequency domain solver have shown, that lowering the line impedance of the parallel-plate waveguide by making the plates broader and their distance to each other smaller improves the response of the inverter. It was chosen not to lower the distance of the plates lower than 1 mm, in order to not compromise the power handling capability of the inverter.

The abs. values of the transmission- and reflection coefficient of the inverter are shown in Fig. 4.26. Between 1.2 GHz and 1.89 GHz, the abs. value of the reflection coefficient is smaller than -10 dB and the abs. value of the transmission coefficient is larger than -0.5 dB. This inverter design is used in the model of a 2.1 m long array of inverter double-loops in Section 4.1.6.

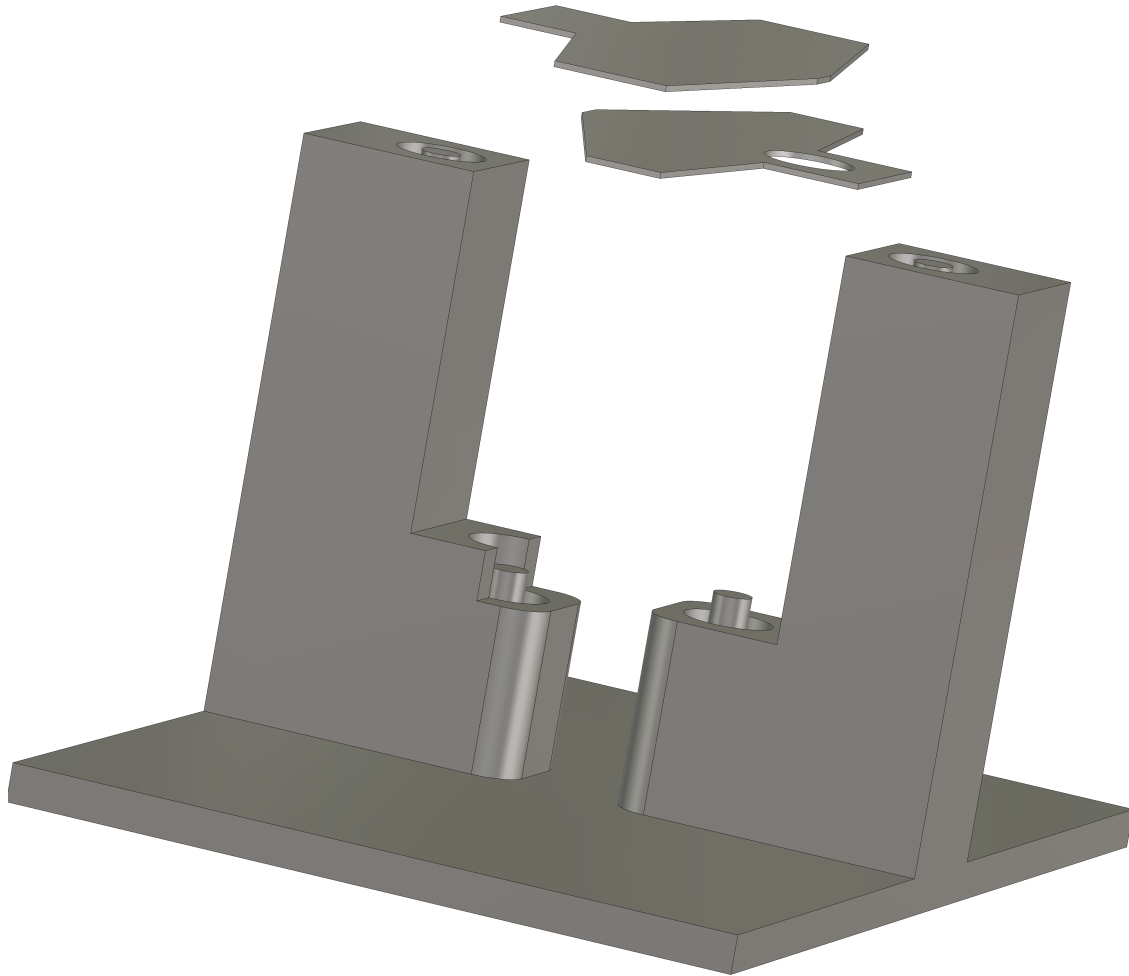


Fig. 4.20: Blown-up view of a double-loop inverter design.

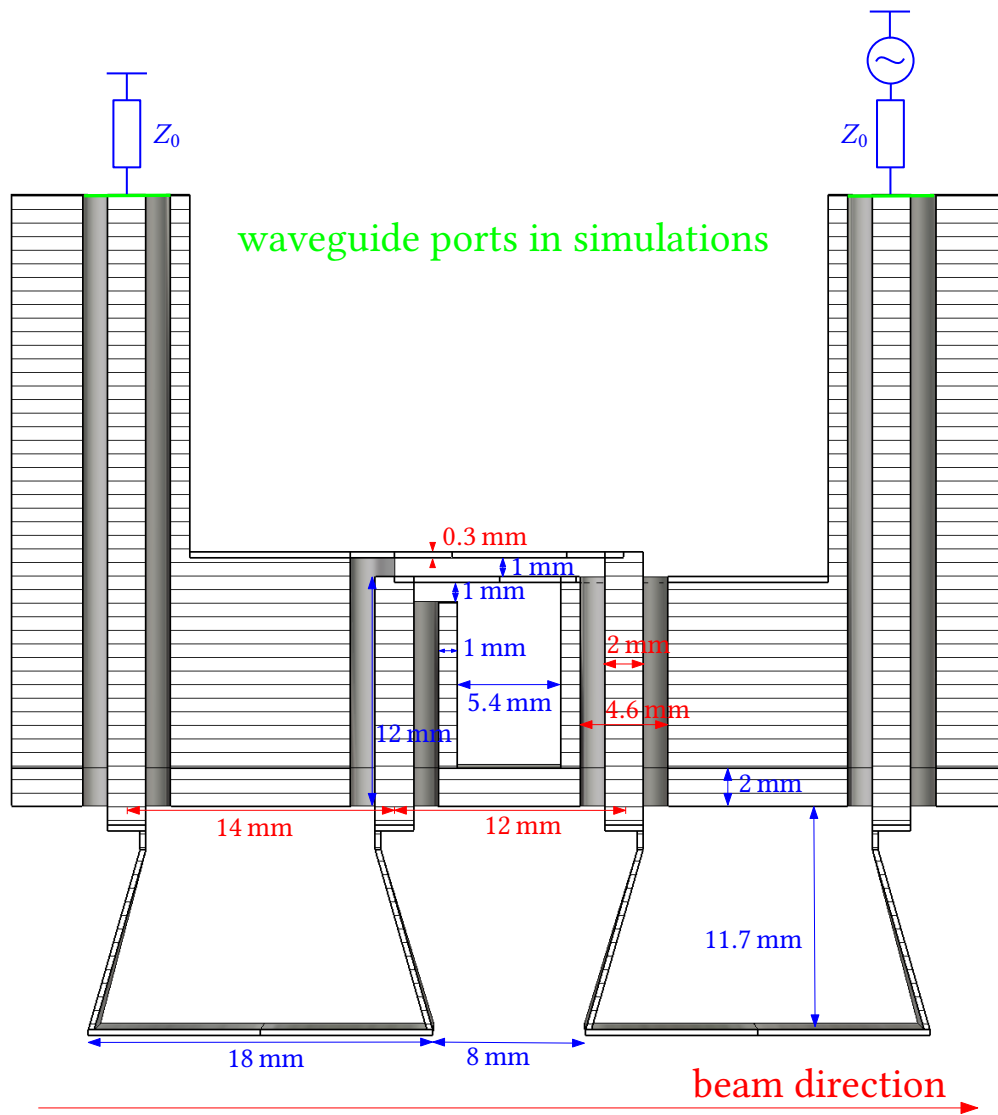


Fig. 4.21: Cross-section view of the double-loop inverter design, including the striplines. The feeding scheme, beam direction and some physical dimensions are illustrated.

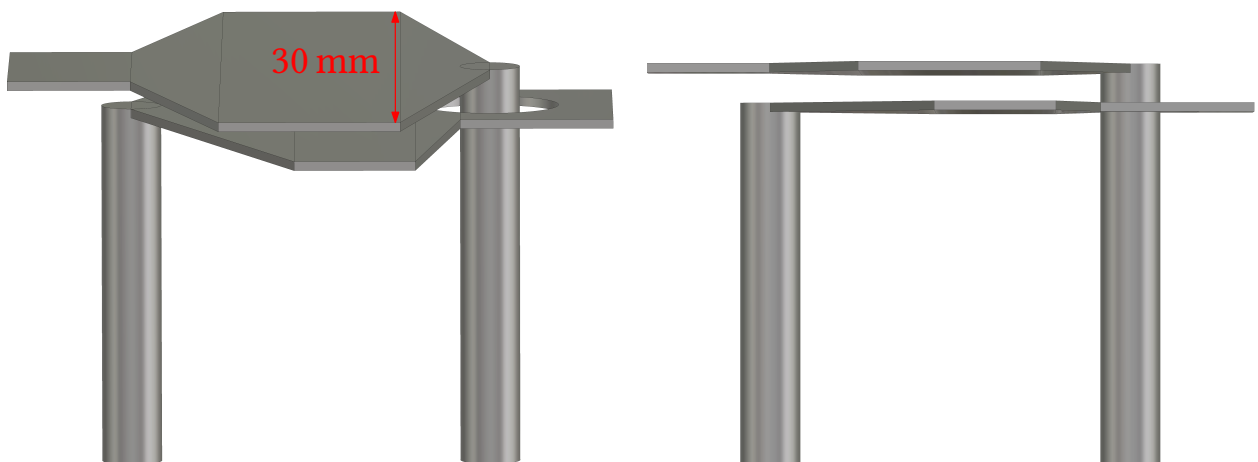


Fig. 4.22: Detailed view of the parallel plate waveguide and the inner conductors of the coaxial-lines of the pulse inverter (left: 3D view, right: 2D side-view).

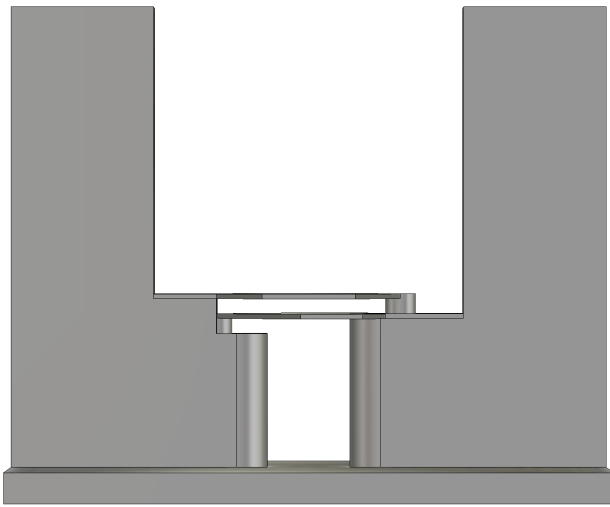


Fig. 4.23: Side-view of a double-loop inverter design.

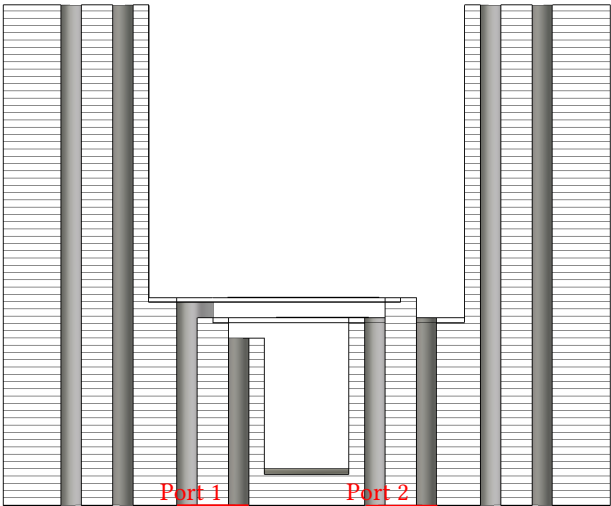


Fig. 4.24: Cross-section side-view of a double-loop inverter design. The ports used for CST simulations of the inverter are shown.

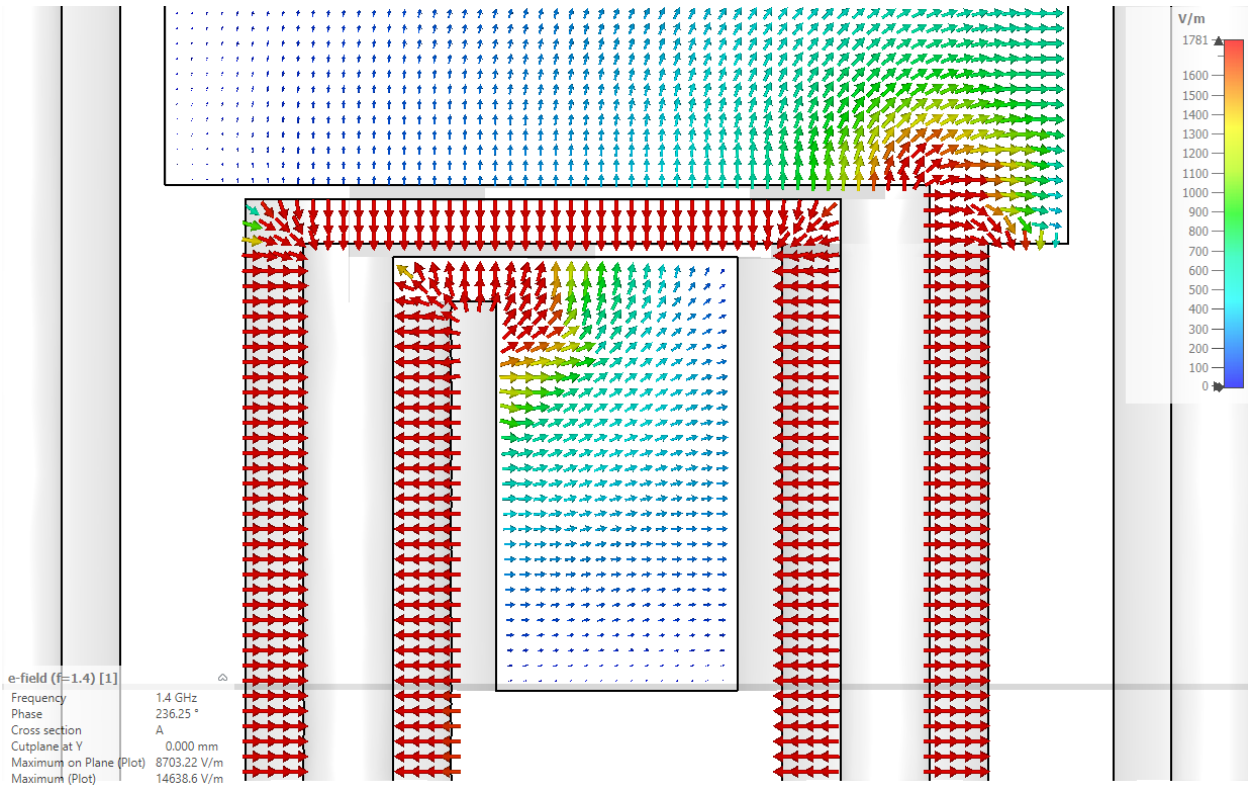


Fig. 4.25: Electric field plot on cross-section of double-loop inverter at 1.4 GHz.

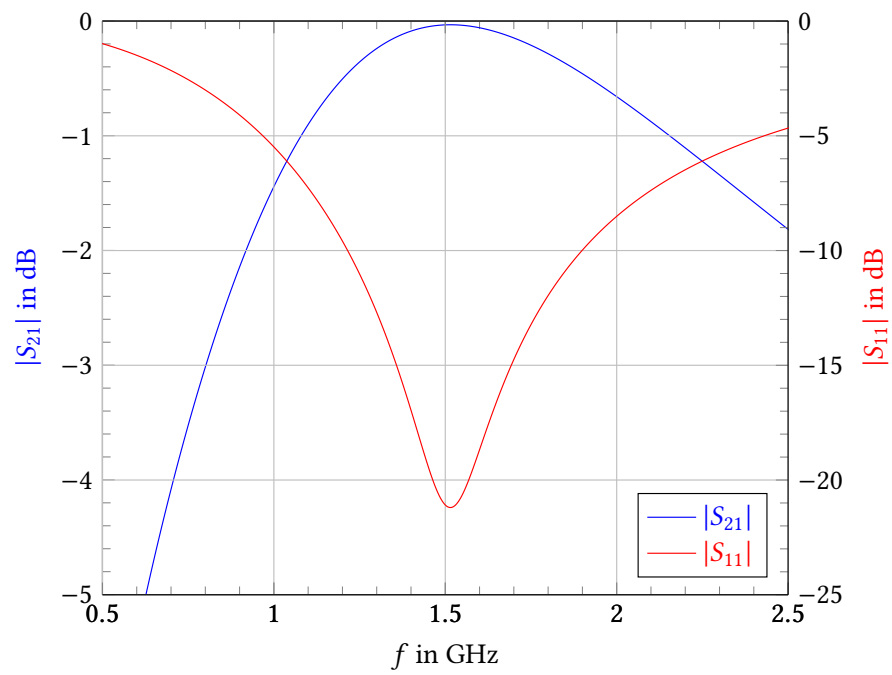


Fig. 4.26: Transmission- and reflection coefficient of double-loop inverter. The port-numbering can be seen in Fig. 4.24.

4.1.5 A Pulse Inverter Design for a Triple-Loop Kicker

A pulse inverter design suitable for a triple-loop kicker is presented. The details of its mechanical design are shown and the electromagnetic transmission behavior investigated. The basic setup and dimensions are equal to those of the design for the double-loop (Section 4.1.4). As necessary for a triple-loop device, the design contains two inverters (refer to Figs. 4.27 to 4.29).

In addition to the transmission- and reflection-coefficient response in Fig. 4.30, the undesired coupling between the ports is shown in Fig. 4.31. The undesired coupling of the ports of the design is below -17 dB, from DC up to well above the operational frequency band of the stochastic cooling system of the AD. The undesired coupling of the ports may be further decreased to below -20 dB by careful design. In the course of this work, no measures are taken against the undesired coupling.

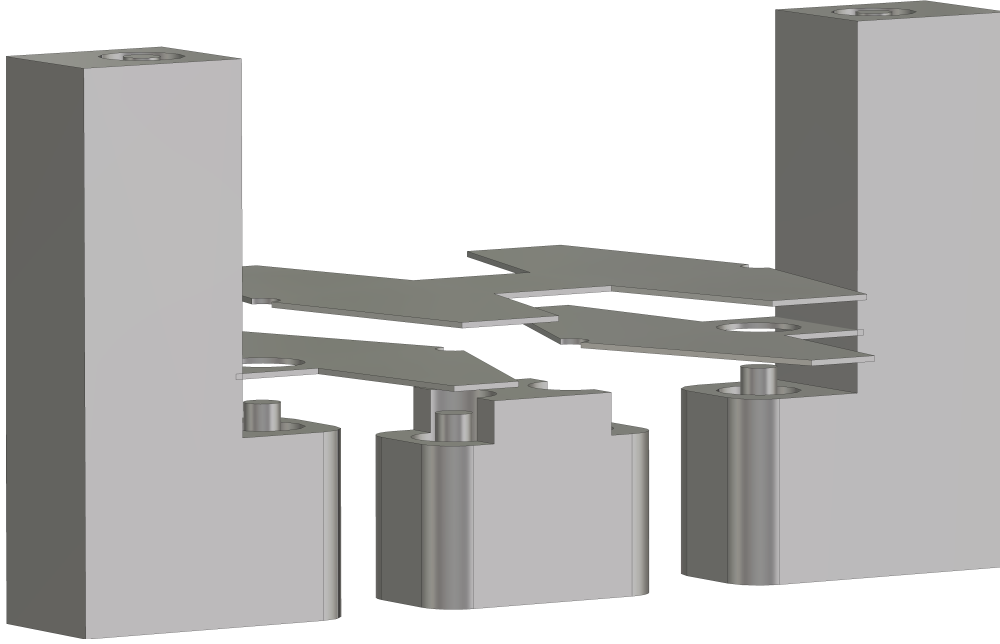


Fig. 4.27: Blown-up view of a triple-loop inverter design.

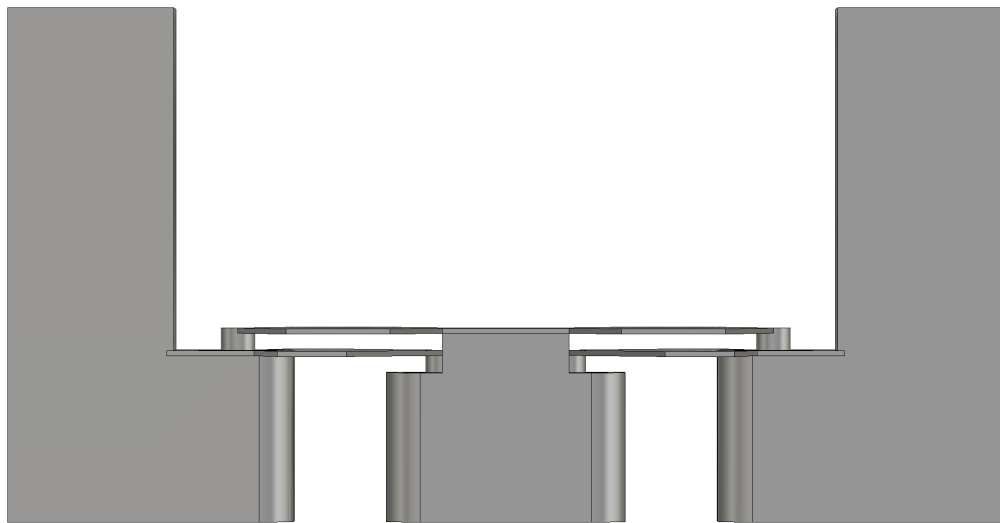


Fig. 4.28: Cross-section side-view of a triple-loop inverter design.

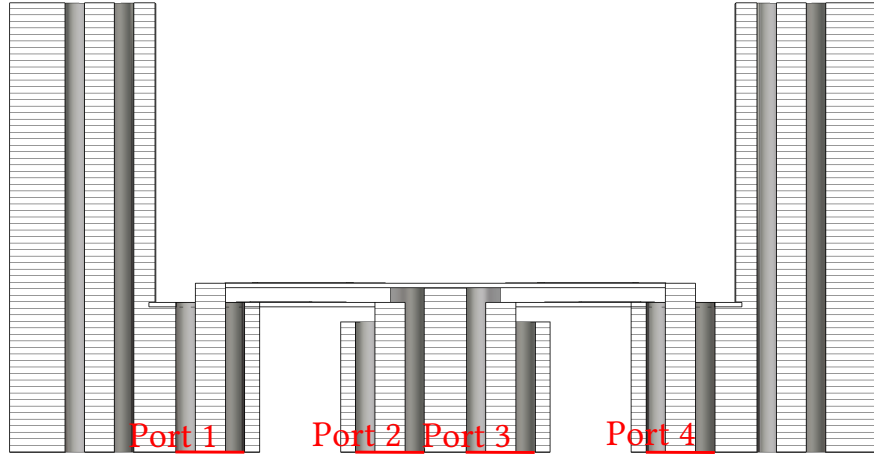


Fig. 4.29: Cross-section side-view of a triple-loop inverter design.

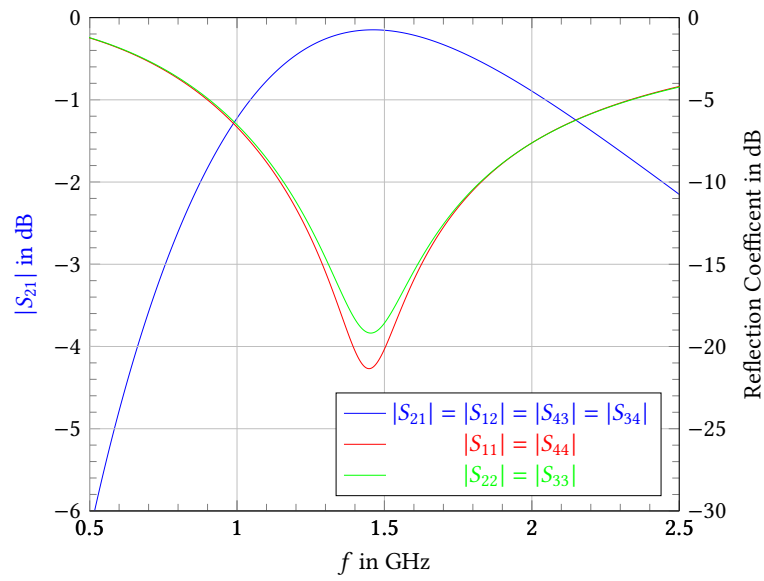


Fig. 4.30: Transmission- and reflection coefficient of triple-loop inverter. The ports can be seen in Fig. 4.29.

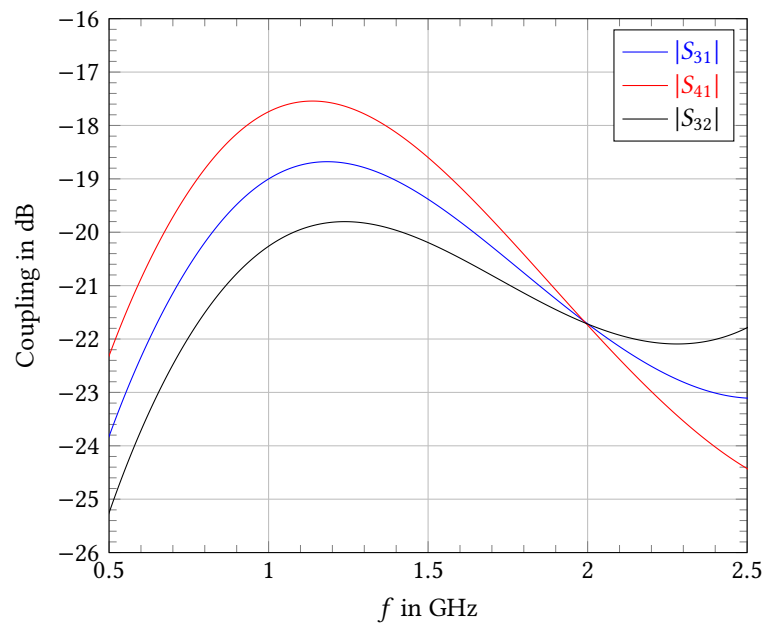


Fig. 4.31: Undesired coupling of the ports of the triple-loop inverter. The ports can be seen in Fig. 4.29.

4.1.6 An Inverter Double-Loop Kicker

In this section, a design of an inverter double-loop kicker array is shown. The design uses an array with an active length of 2.1 m and the adjacent beampipes of the KHM kicker tank (refer to Figs. 4.32 and 4.33 for detailed views of sections of the kicker). Aperture dimensions of $e_{2ed} = 80.5$ mm and a beampipe height of 92.9 mm are used. A total of 40 inverter double-loops are contained in the 2.1 m active length. The abs. values of the total longitudinal and transverse kick factors are shown in Figs. 4.34 and 4.35. Zoomed-in versions of the plots are shown in Figs. 4.36 and 4.37. The nonlinear phase of the longitudinal and transverse kick factors are provided in Figs. 4.48 to 4.51. All results are provided for the case, where ferrites are used in the kicker volume and for the case where no ferrites are used.

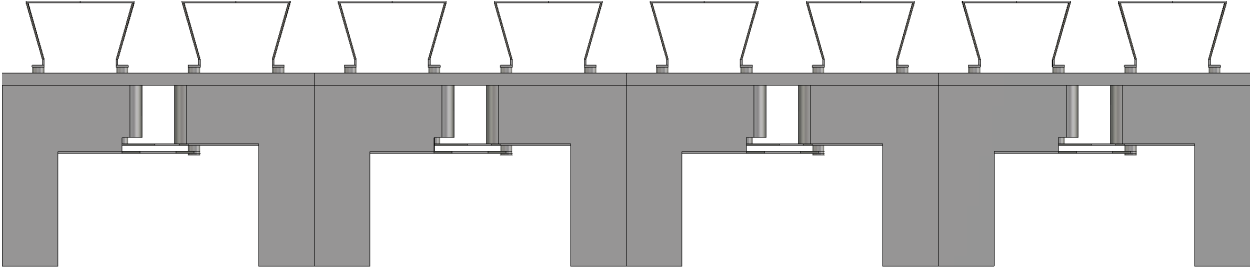


Fig. 4.32: Side-view of a section of an inverter double-loop kicker.

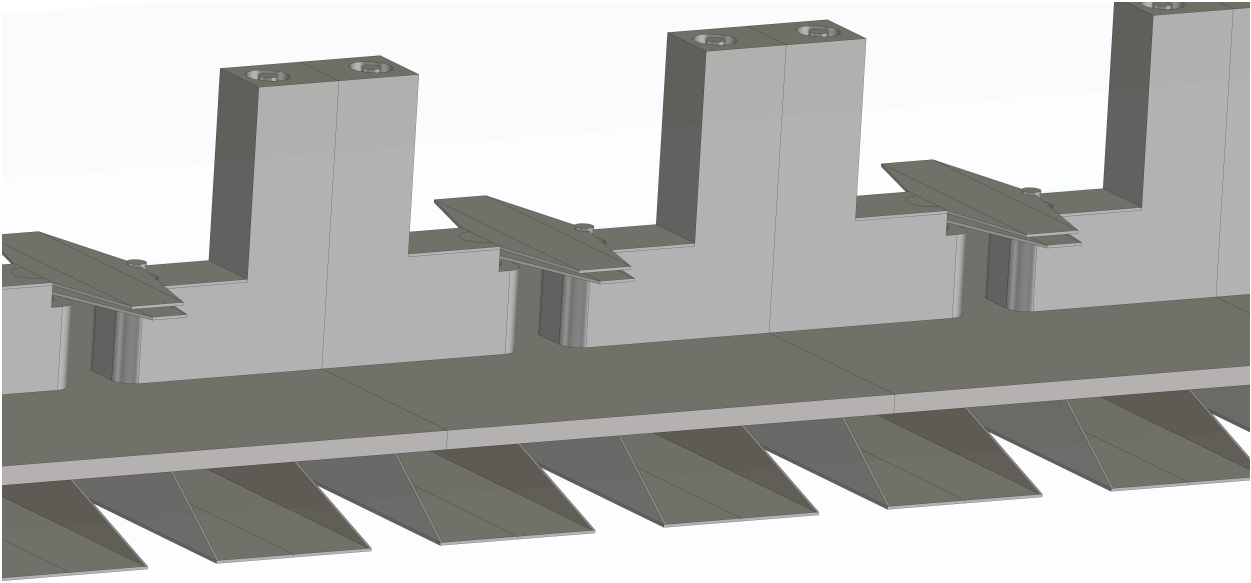


Fig. 4.33: 3D-view of a section of the inverter double-loop kicker geometry.

As seen in the longitudinal response (Figs. 4.36 and 4.37), the zero at 800 MHz is successfully suppressed, compared to the response of the classic double-loop. Compared to the classic double-loop design, the upper frequency limit is reduced by a small amount, as the zero at approx. 2.3 GHz in the classic double-loop response is now located at approx. 2.1 GHz. Comparing Fig. 4.35 to Fig. 3.4, it can be seen that the inverter double-loop design suppresses the large DC peak in the transverse kick factor.

The suppression of the zero at 800 MHz can also be observed in the nonlinear phase of the kick factor, as there is no phase jump at 800 MHz, compared to the classic double-loop response. The nonlinear phase is smaller than $\pm 5^\circ$ throughout most of the relevant frequency band (Figs. 4.50 and 4.51). A comparison of this inverter double-loop kicker design to all other designs of this work is given in Section 4.3.

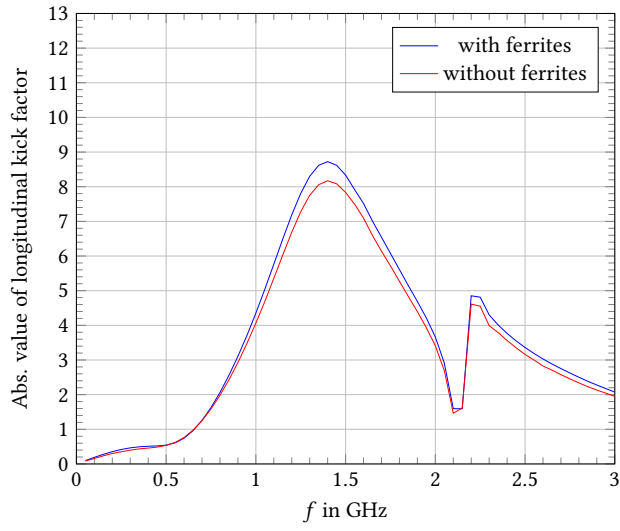


Fig. 4.34: Abs. value of longitudinal kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$.

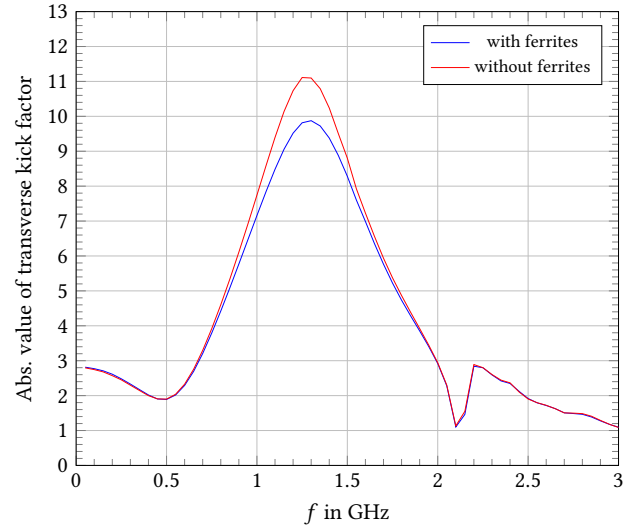


Fig. 4.35: Abs. value of transverse kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$.

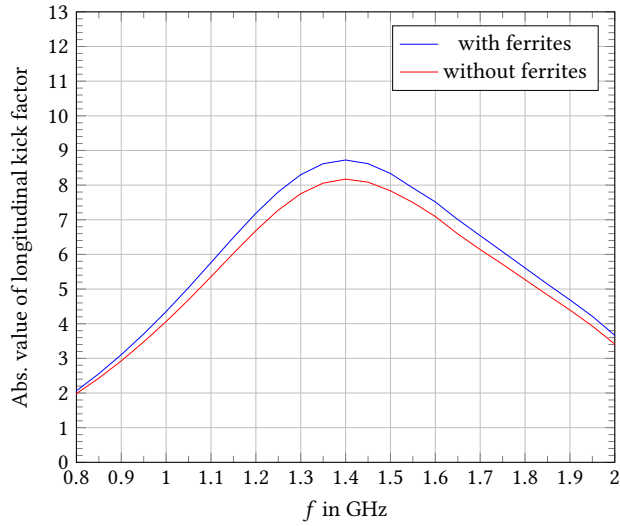


Fig. 4.36: Abs. value of longitudinal kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$.

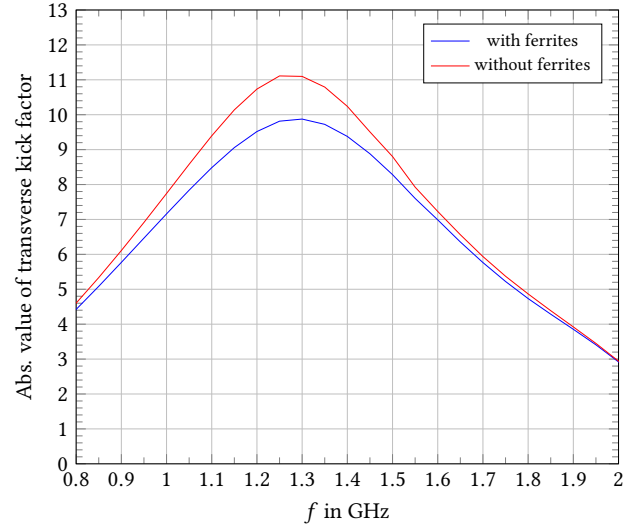


Fig. 4.37: Abs. value of transverse kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$.

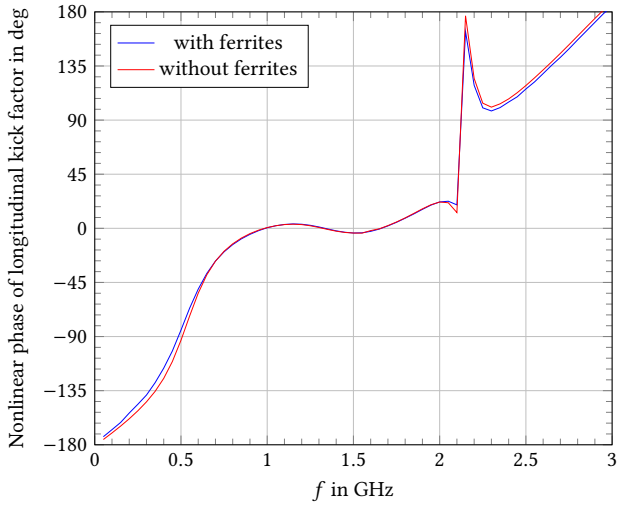


Fig. 4.38: Nonlinear phase of longitudinal kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

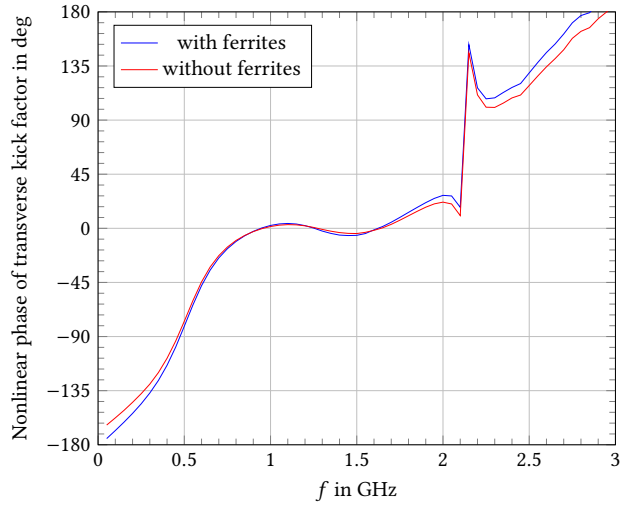


Fig. 4.39: Nonlinear phase of transverse kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

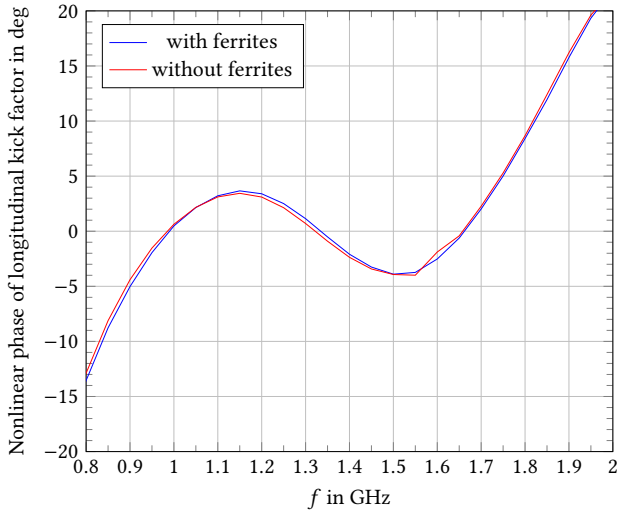


Fig. 4.40: Nonlinear phase of longitudinal kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

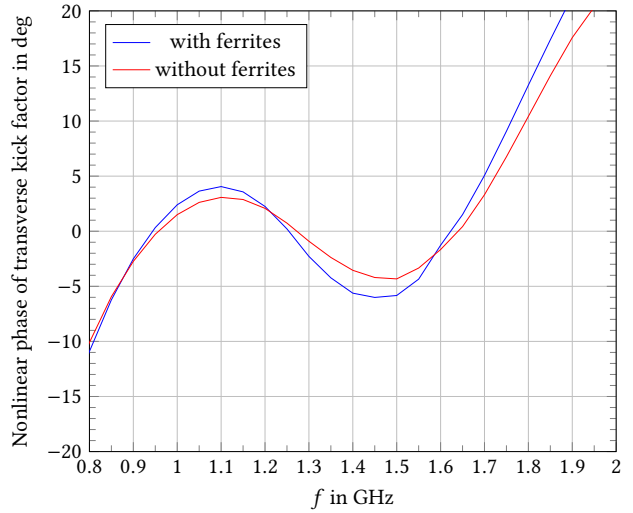


Fig. 4.41: Nonlinear phase of transverse kick factor of an inverter double-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

4.1.7 An Inverter Triple-Loop Kicker

In this section, a design of an inverter triple-loop kicker array is shown. Figs. 4.42 and 4.43 show detailed views of sections of the kicker. The design contains a kicker array with an active length of 2.1 m and the adjacent beampipes of the KHM kicker tank. Aperture dimensions of $e_{2ed} = 80.5$ mm and a beampipe height of 92.9 mm are used. A total of 30 inverter triple-loops are contained in the 2.1 m active length. The abs. values of the total longitudinal and transverse kick factors are shown in Fig. 4.44 and Fig. 4.45. Zoomed-in versions of the plots are shown in Figs. 4.46 and 4.47. The nonlinear phase of the longitudinal and transverse kick factors are provided in Figs. 4.48 to 4.51.

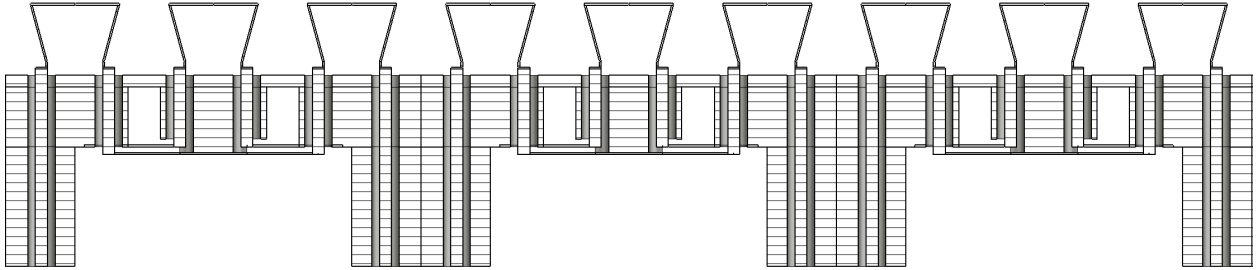


Fig. 4.42: Side view of a section in the inverter triple-loop kicker.

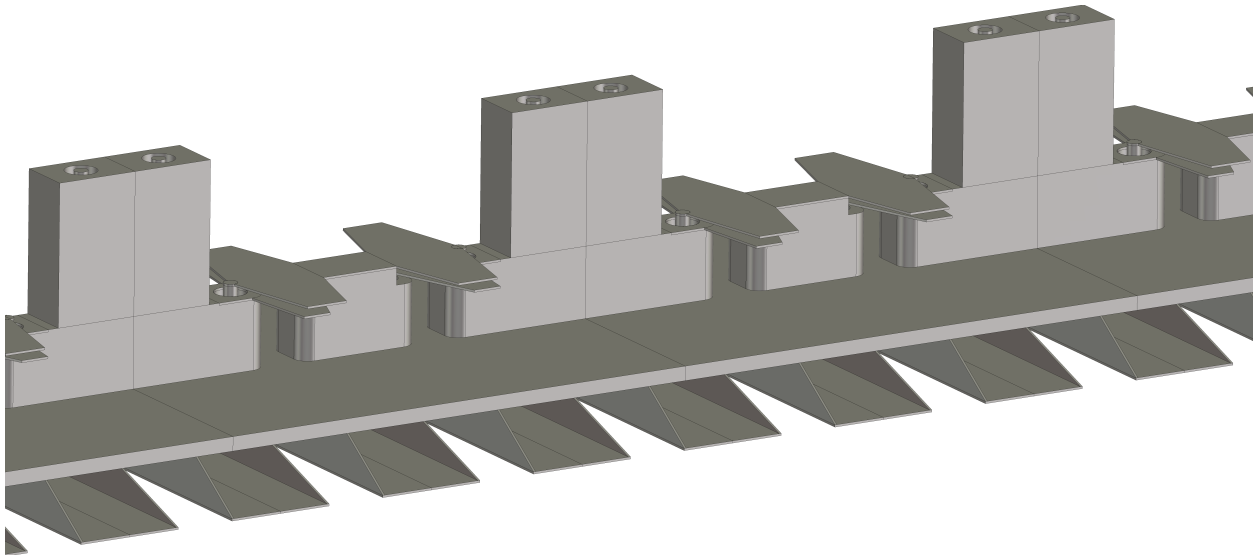


Fig. 4.43: Detailed 3D-view of a section in the inverter triple-loop kicker.

It can be seen that the triple-loop characteristic zeros are successfully suppressed, such that the zeros in the response are almost equal to the zeros in the response of the classic double-loop design (refer to Figs. 3.3 and 3.4).⁵

The results demonstrate, that an inverter triple-loop design provides higher peak values than classic double-loops while having comparable bandwidth and this without needing significantly more feed-ports compared to inverter double-loops or slot-line designs. A comparison of this inverter triple-loop kicker design to all other designs of this work is given in Section 4.3.

⁵Also compare the simulation results to the plots of the analytical models of triple-loops shown in Fig. 4.2.

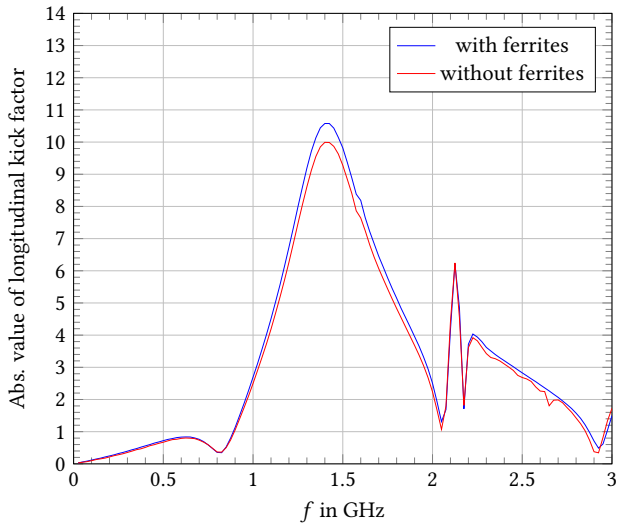


Fig. 4.44: Abs. value of longitudinal kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$.

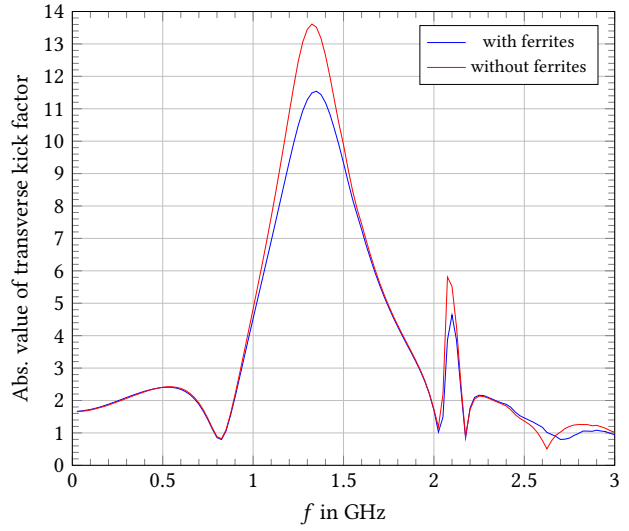


Fig. 4.45: Abs. value of transverse kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$.

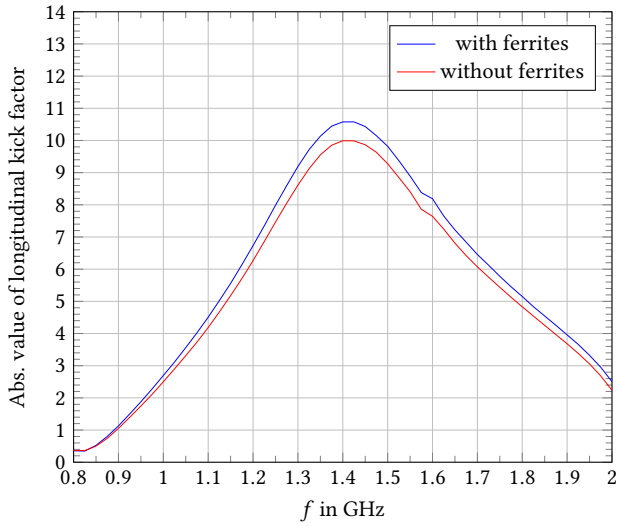


Fig. 4.46: Abs. value of longitudinal kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$.

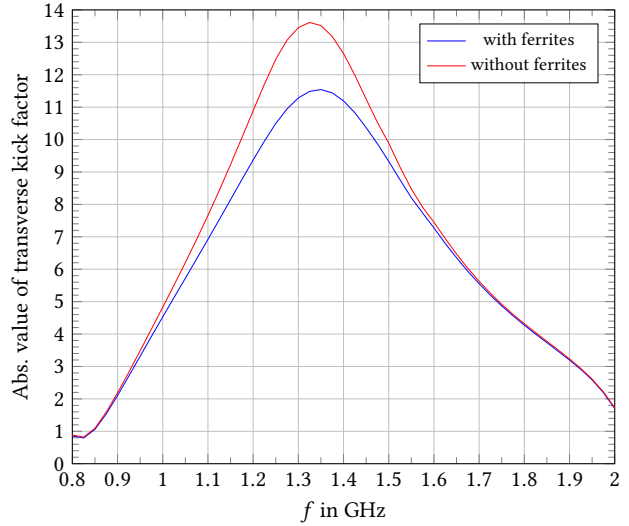


Fig. 4.47: Abs. value of transverse kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$.

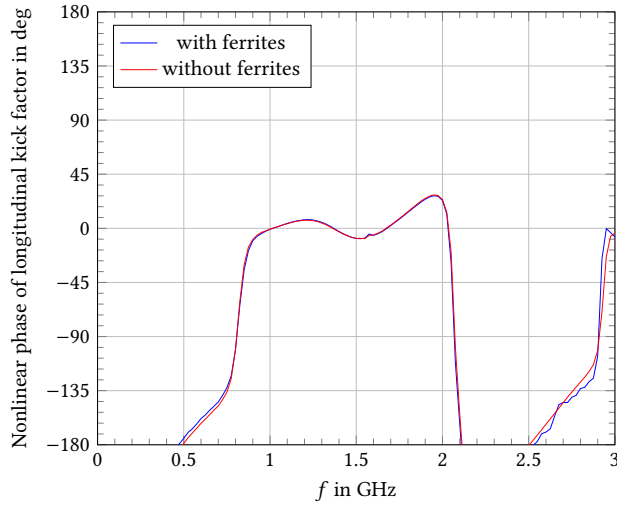


Fig. 4.48: Nonlinear phase of longitudinal kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

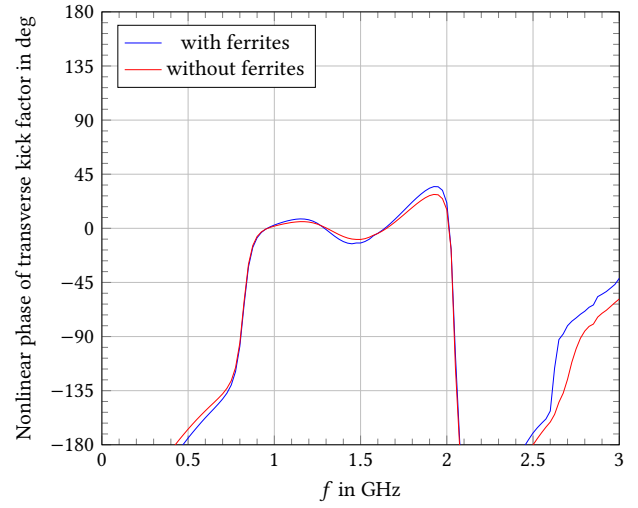


Fig. 4.49: Nonlinear phase of transverse kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

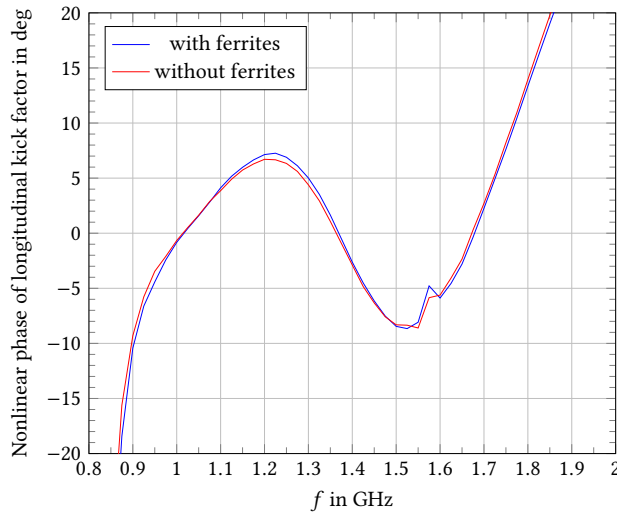


Fig. 4.50: Nonlinear phase of longitudinal kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

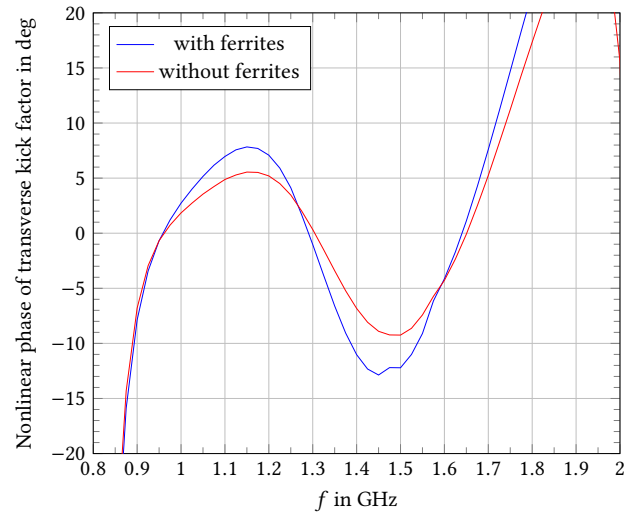


Fig. 4.51: Nonlinear phase of transverse kick factor of an inverter triple-loop array of 2.1 m length for $\beta = 0.96724$. Nonlinear phase computed with respect to a linear least squares fit in the 0.9 GHz to 1.8 GHz band.

4.2 A Slotline Kicker Design

In this section, a slotline kicker based on a proposal by Thorndahl [45] is investigated. The structure is based on the TSWS by Caspers [19] (refer to Section 2.3). A design of a 2.1 m long kicker with aperture dimensions of $e_{2ed} = 80.5$ mm and a beampipe height of 92.9 mm is shown in Figs. 4.52 to 4.54. Beampipes of the current KHM kicker tank are used. As seen in Figs. 4.53 and 4.54, the slots are coupled to by striplines, which are formed by a piece of conductor and the wall of the individual slot compartments. The striplines are electrically shorted to the slots.

By coaxial feedthroughs, the slots can be fed with power without power division circuitry inside the vacuum tank. Why this would be desired was discussed in Section 2.4. As the slots are not terminated by an RF load, the full power fed to the slots is reflected, except for power which leaves the beampipes due to the excitation of waveguide modes and power that is lost due to ohmic losses in the slot compartments. The reflected power needs to be absorbed outside of the kicker tank. The power amplifiers which feed the kicker need to be able to handle the reverse power, or circulators have to be used.

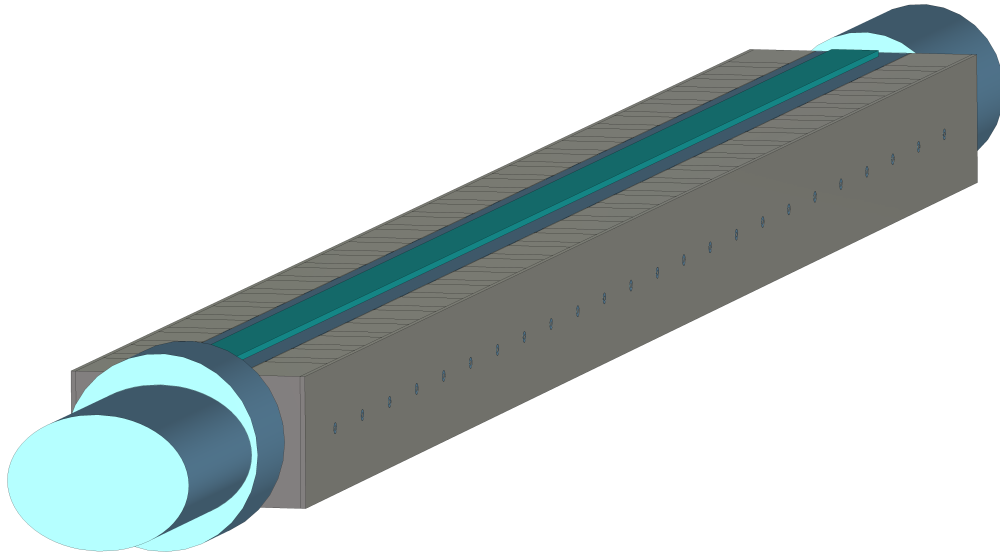


Fig. 4.52: 3D view of the CST model of the slotline kicker array. The active length is 2.1 m and KHM beampipes are connected to the kicker.

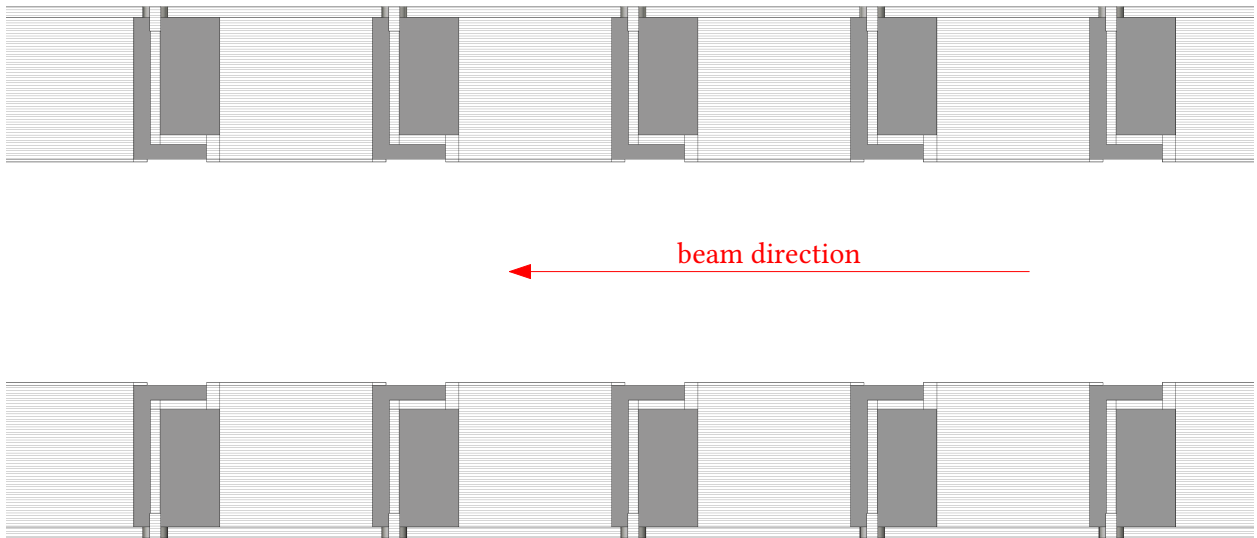


Fig. 4.53: 2D view of a side-cut through the slotline kicker.

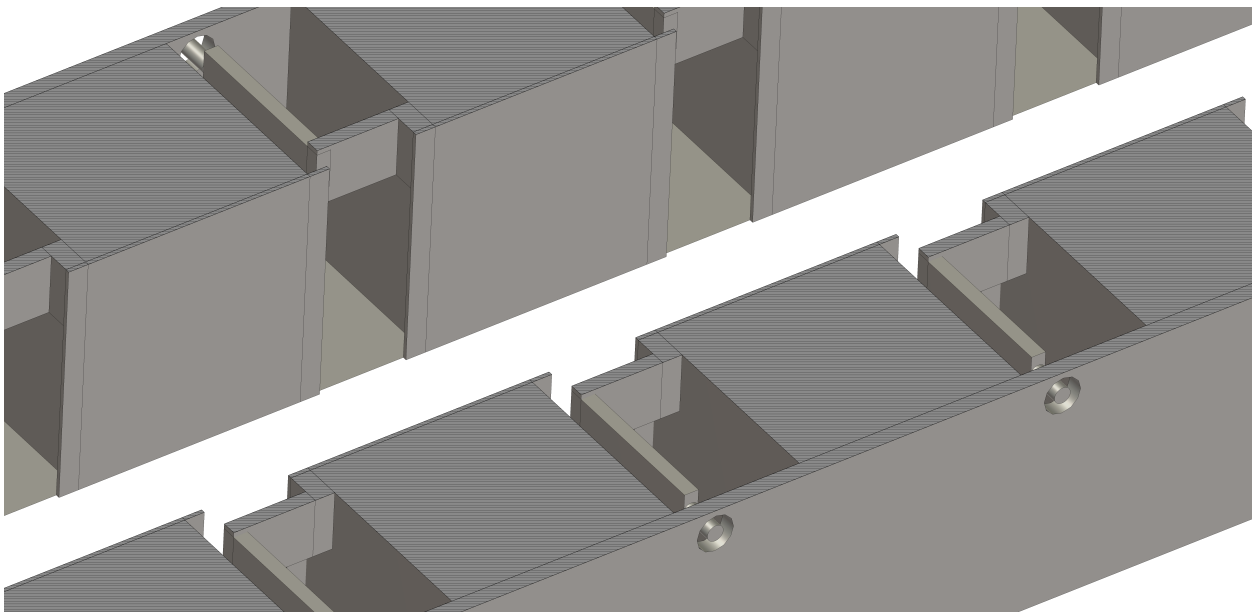


Fig. 4.54: 3D-view of a cut through the slotline kicker.

The structure has been optimized to achieve peak values of the kick factor, that are comparable to those of the currently used double-loop design (Chapter 3) without using a large number of slots. The simulations were performed on a kicker of 2.1 m active length. The total number of elements contained in the active length has been swept from 20 to 60. The abs. values of the longitudinal and transverse kick factors are shown in Figs. 4.55 to 4.58 and the deviation of the longitudinal and transverse responses from linear phase are provided in Figs. 4.59 to 4.62.

It can be seen, that for 32 or more unit cells, the longitudinal and transverse response get competitive to the inverter double-loop design, both in the longitudinal and transverse response and both in magnitude and phase. The phase response does not deteriorate, if the slots are placed closely to each other. A comparison of this slotline kicker design to all other designs of this work is given in Section 4.3.

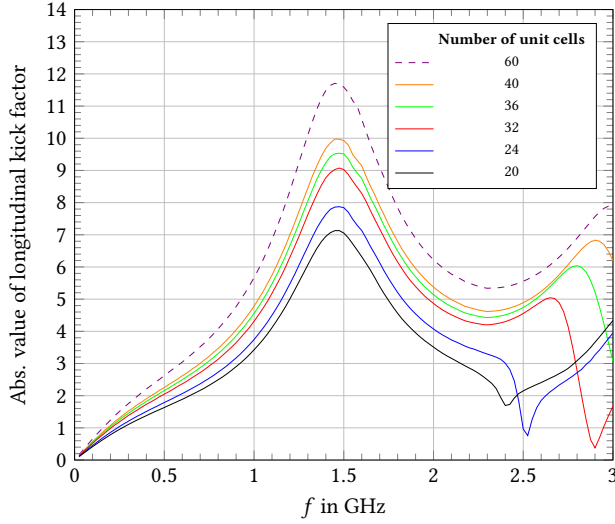


Fig. 4.55: Abs. value of longitudinal kick factor of a slotline kicker.

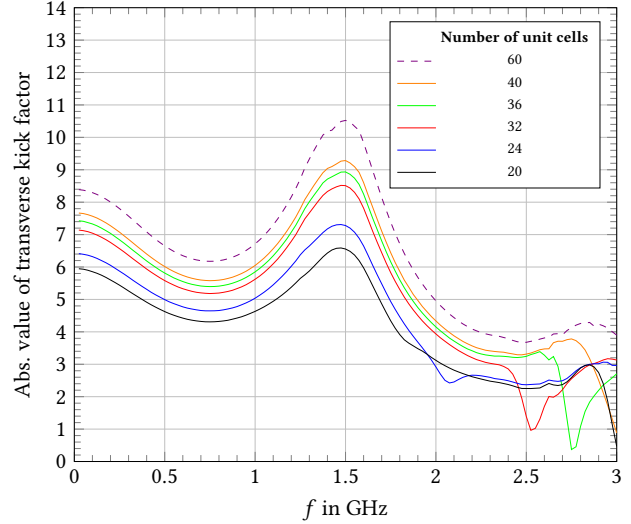


Fig. 4.56: Abs. value of transverse kick factor of a slotline kicker.

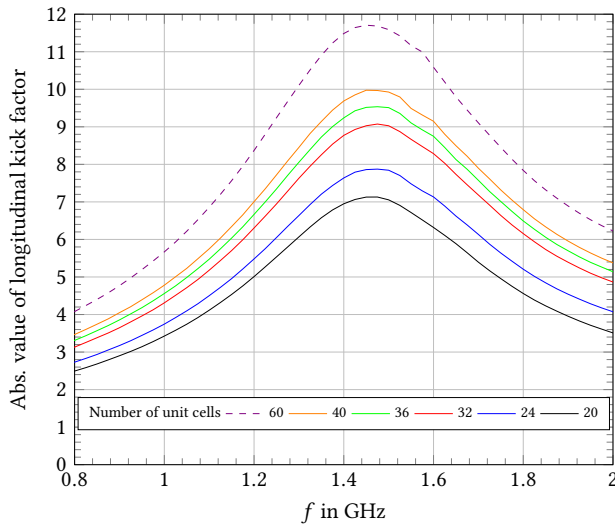


Fig. 4.57: Abs. value of longitudinal kick factor of a slotline kicker.

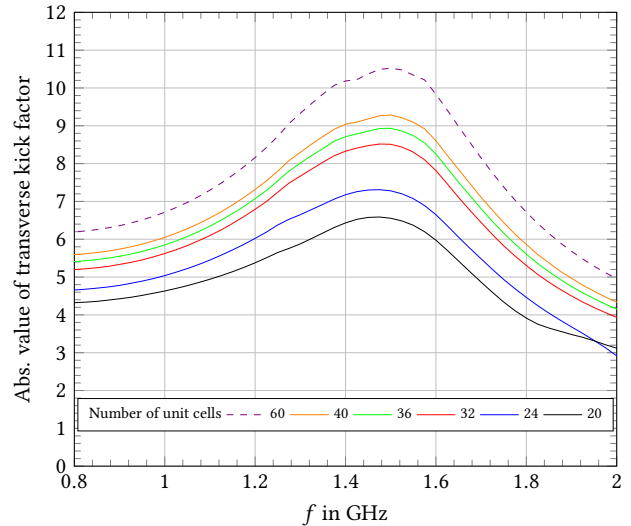


Fig. 4.58: Abs. value of transverse kick factor of a slotline kicker.

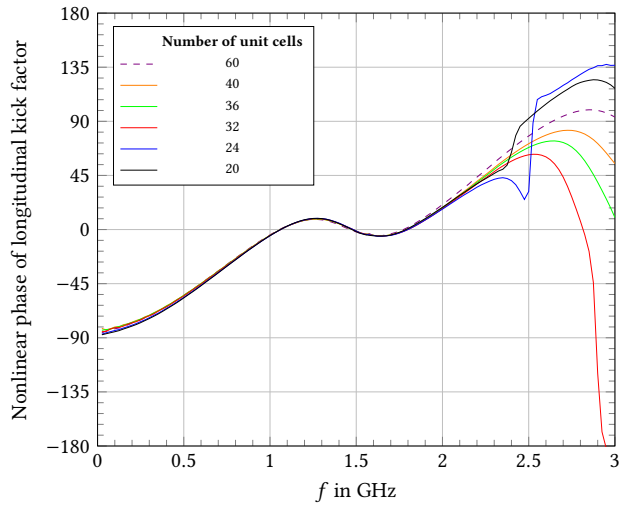


Fig. 4.59: Nonlinear phase of longitudinal kick factor of a slotline kicker.

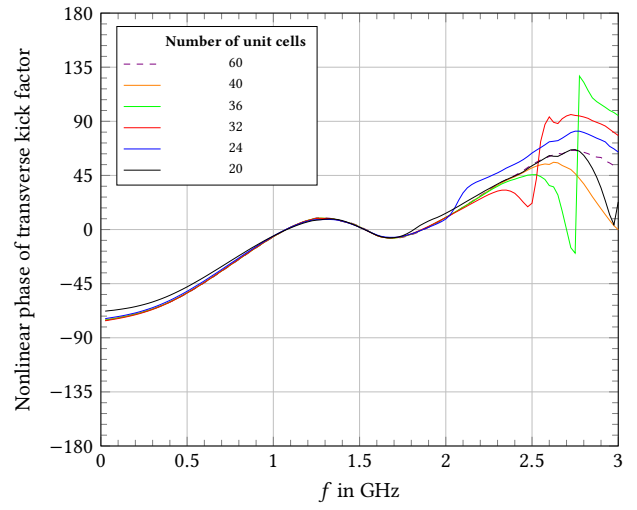


Fig. 4.60: Nonlinear phase of transverse kick factor of a slotline kicker.

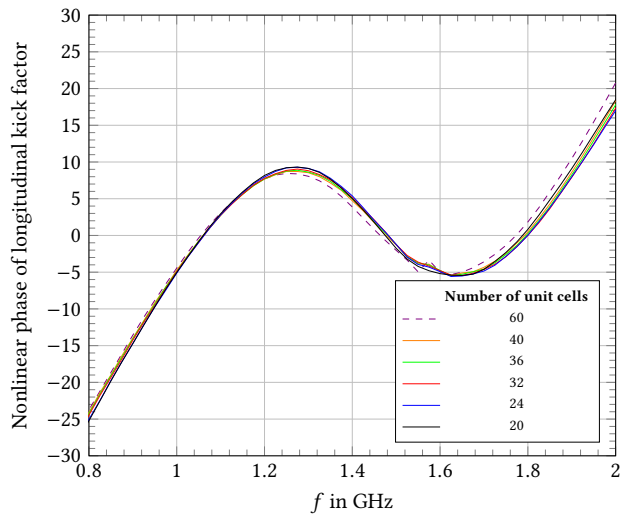


Fig. 4.61: Nonlinear phase of longitudinal kick factor of a slotline kicker.

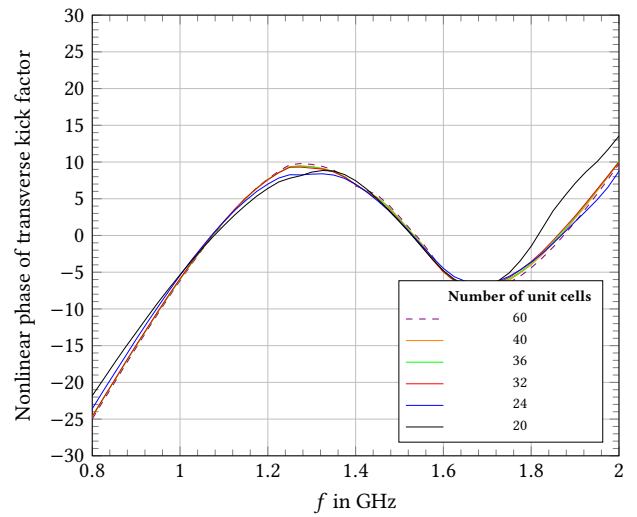


Fig. 4.62: Nonlinear phase of transverse kick factor of a slotline kicker.

4.3 Comparison of the Designs

In this section, the designs presented in Chapter 4 are compared to each other, and to the classic double-loop kicker design that is currently in use in the AD (refer to Chapter 3). The kicker designs are compared to each other with respect to their longitudinal and transverse kick factor at $\beta = 0.96724$. Comparisons of the longitudinal and transverse kick factors of all the introduced double- and triple-loop kicker designs are found in Figs. 4.63 and 4.64. Note that in the different designs, a different number of individual kick elements fit in the total active length of 2.1 m. For example, the inverter double-loop kicker design uses 40 individual inverter double-loops. All designs have the same aperture dimensions of $e_{2ed} = 80.5$ mm and a beam pipe height of 92.9 mm.

In the longitudinal response (Fig. 4.63) it can be seen, that the inverter triple-loop design has a larger kick factor than the currently used design, for frequencies below approx. 1.8 GHz. The inverter double-loop design has improved bandwidth and peak value, while the inverter triple-loop has slightly lower bandwidth than the currently used double-loop design. In peak value, the inverter triple-loop design has an improvement of more than 20% compared to the currently used double-loop design.

In the transverse response (Fig. 4.64), the inverter double-loop design has the same upper frequency limit like the currently used design, while having larger kick factor at lower frequencies down to 800 MHz. This is due to the suppression of the zero at 800 MHz. The inverter triple-loop design has significantly more peak value than the currently used double-loop design, has also a zero at approx. 800 MHz but has a smaller kick factor for frequencies above approx. 1.7 GHz.

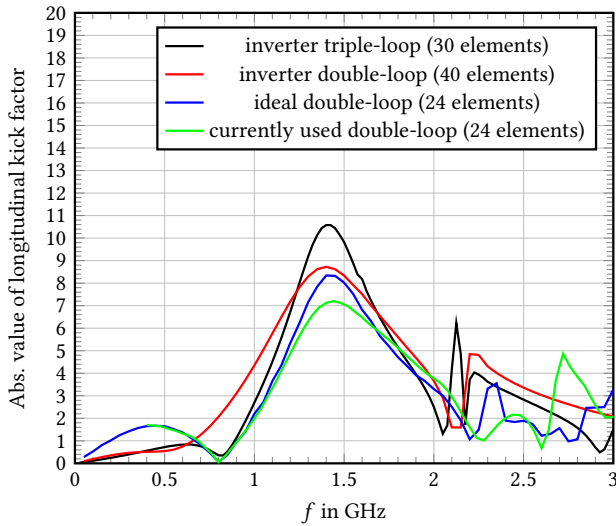


Fig. 4.63: Abs. value of long. kick factor at $\beta = 0.96724$. Comparison of the inverter double-loop and inverter triple-loop kickers from Sections 4.1.6 and 4.1.7, to the ideal double-loop kicker from Section 4.1.2 and the response of the currently used kickers from Chapter 3.

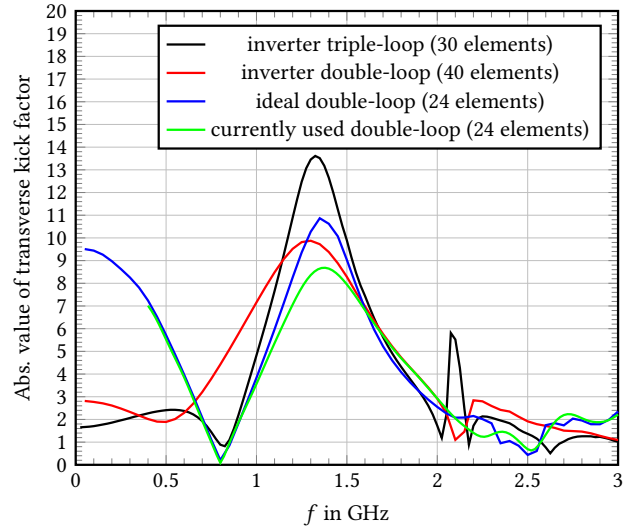


Fig. 4.64: Abs. value of transv. kick factor at $\beta = 0.96724$. Comparison of the inverter double-loop and inverter triple-loop kickers from Sections 4.1.6 and 4.1.7, to the ideal double-loop kicker from Section 4.1.2 and the response of the currently used kickers from Chapter 3.

A comparison of the inverter multi-loops to the slotline kicker design from Section 4.2 is shown in Figs. 4.65 and 4.66. It can be seen, that the longitudinal response of the slotline kicker design is competitive to the inverter multi-loop designs, even for a low number of slots (equal to, or smaller than 40). For a slot-number of 40 or lower, the peak value of the transverse response of the slotline kicker design, is approx. equal to the peak value of the inverter double-loop and smaller than the peak value of the inverter triple-loop. Here, the bandwidth of the slotline design is approx. equal to the bandwidth of the inverter double-loop design and larger than the bandwidth of the inverter triple-loop design.

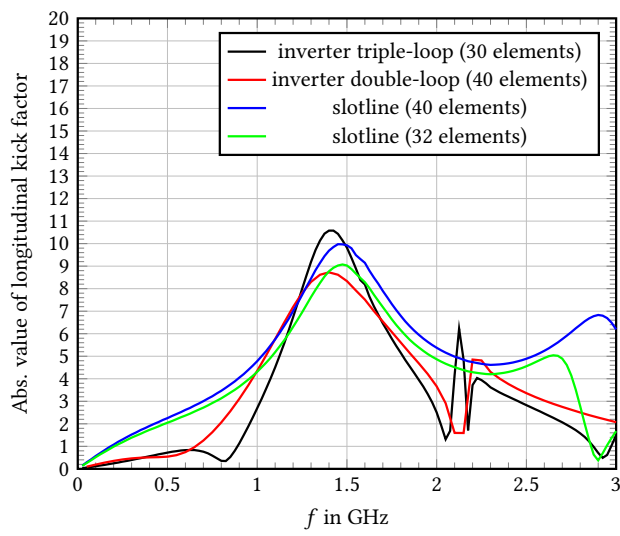


Fig. 4.65: Abs. value of long. kick factor at $\beta = 0.96724$. Comparison of the inverter double-loop and inverter triple-loop kickers from Sections 4.1.6 and 4.1.7, to slotline kickers from Section 4.2.

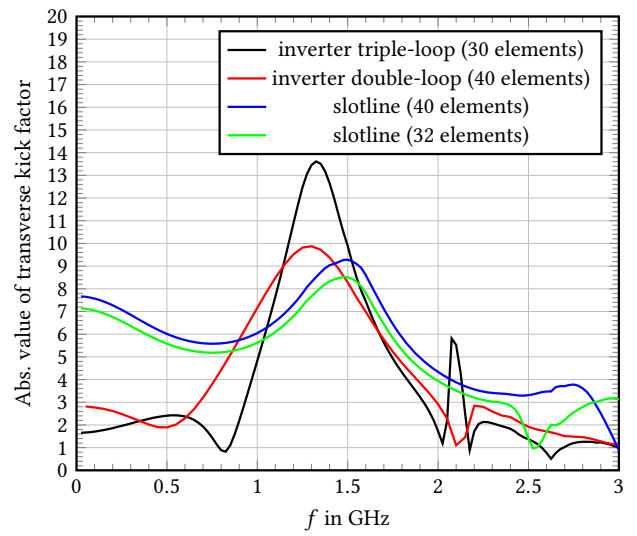


Fig. 4.66: Abs. value of transv. kick factor at $\beta = 0.96724$. Comparison of the inverter double-loop and inverter triple-loop kickers from Sections 4.1.6 and 4.1.7, to slotline kickers from Section 4.2.

5 Conclusion & Acknowledgements

In this work, the pickups and kickers currently used in the stochastic cooling system of the AD at CERN were analyzed by means of numerical simulations using CST. The performance at the two momentum plateaus used in the momentum program of the AD was computed, taking into account the signal combination and power splitting networks that are used today in the AD. Furthermore it has been derived, that the complete response of the pickup- and kicker arrays can be analytically approximated by multiplying the well known response functions of striplines by an intra-loop array factor and a combination/splitting array factor.

A novel design based on striplines coupled by RF pulse inverters was introduced and investigated. 2.1 m long kicker arrays of the novel devices have been designed. It has been shown, that by the use of the inverter triple-loop design, a kicker response with improved peak value compared to classic double-loops is obtained. The bandwidth performance is reduced by a small amount, compared to the currently used double-loops. Nevertheless, its bandwidth is larger than for classic triple-loops, which is due to the inverter suppressing the triple-loop characteristic zeros in the response. It was demonstrated, that the inverter double-loop topology suppresses the 800 MHz zero in the kicker response.

Finally, a design of a 2.1 m long slotline kicker array has been presented. By optimization of the structure it was shown, that the kicker response is competitive to classic- and inverter double-loops in terms of peak value and bandwidth of the response. This has been achieved with a design that does not employ a very large number of individual kick elements and thus does not require power splitting inside the vacuum tank of the kicker.

Acknowledgements

First and foremost, I would like to express my sincere gratitude to my supervisor, Wolfgang Hofle, for his continuous support, valuable guidance, and encouragement throughout this thesis. He has helped me see the broader context of my work beyond the specific scope of this project. I am deeply grateful for every minute he spent explaining topics in RF and accelerator physics and for the opportunity to visit and work on the installations at CERN under his guidance. Learning from him has been a great honor and will have an immeasurable impact on my future career.

I would also like to thank Lars Thorndahl, who designed a large part of the AD stochastic cooling pickups and kickers. Learning from such an experienced engineer and physicist has been a true privilege. The guidance of Bernd Breitzkreutz and Manfred Wendt was a great help during the beginning of the internship. I am also grateful to Stéphane Rey for generously taking the time to show me the inner workings of the AD and for sharing his insights along the way. Many discussions with Leonard Thiele and Stefan Wunderlich (GSI) helped me a lot in improving my skills in numerical simulations using CST.

I would like to thank Prof. Eibert from the Technical University of Munich (TUM), whose lectures laid the foundation on which both this thesis and my Technical Studentship at CERN were built. Last but not least, I would like to thank the late Fritz Caspers for all the time he spent explaining RF and accelerator physics topics to me. His historical remarks were always amusing and added a unique charm to his explanations. I think it is safe to say he had an immeasurable impact on generations of engineers and physicists in the particle accelerator community.

A Striplines

In this Appendix, the formulas in Tab. 2.1 and Tab. 2.2, namely the transfer impedance, kick factor and shunt impedance of striplines are derived.

To this end, the stripline device is first analyzed in pickup operation (refer to Fig. 2.3). Then the kick factor and shunt impedance are calculated from the transverse and longitudinal transfer impedances using (2.21) and (2.22). Note that the pickup signals are extracted upstream and that the kicker is fed downstream.

Let us assume a beam current density and its corresponding beam current

$$\mathbf{j}(\Delta x, \Delta y, z, t) = qv_{\text{beam}}\lambda(z - v_{\text{beam}}t)\delta(x - \Delta x)\delta(y - \Delta y)\mathbf{e}_z \quad (\text{A.1})$$

$$I_b = i(z, t) = qv_{\text{beam}}\lambda(z - v_{\text{beam}}t). \quad (\text{A.2})$$

The image i_m of the beam current flows on the inner surface of the beam pipe at the same velocity as the beam current. Assuming that the beam velocity is close to the speed of light, the fields stemming from the beam current are contracted in space. This means, that the image current on the beam pipe has the same longitudinal profile $\lambda(z - v_{\text{beam}}t)$ as the beam current. Note that the image charge of a stationary charge q inside a beam pipe would be distributed across the full length of the beam pipe up to infinity. As by charge conservation, the total charge of the image is $-q$ and thus the image current must be $i_m(z, t) = -i(z, t)$.

Now comes a crucial step in the derivation process. As the striplines of width w do not extend over the whole circumference of the beam pipe, a fraction α_B of the image current intercepts the lower transmission line and the termination impedance presented by the 180° Hybrid. The same happens on the upper transmission line at a different fraction α_A . As both the transmission lines and the Hybrid's input ports are matched to Z_{line} , the current $\alpha_B i_m$ sees an impedance of $Z_{\text{line}}/2$ when it intercepts the point $z = a$. A voltage waveform equal to $\alpha_B i_m Z_{\text{line}}/2$ can be measured for $V_B(t)$ and a forward traveling voltage wave is excited on the transmission line. When the image current exits the pickup at $z = b$, it is pulled from the parallel circuit of the matched transmission line and the termination impedance and thus excites a backwards traveling voltage wave of opposite polarity. No signal appears at the port at $z = b$, as the forward and backwards traveling voltage waves cancel there.¹

The circuit in Fig. A.1 is equivalent to the just described process. Such circuit models, equivalent to stripline kickers and pickups, are used in [46, p. 141], [14, p. 24] and [34] for example.

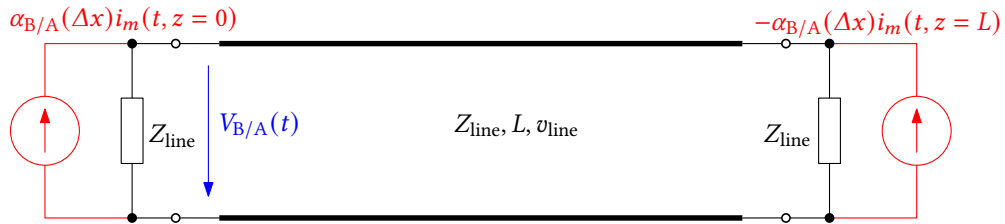


Fig. A.1: Equivalent circuit for the process of the image current intercepting and exiting the stripline.

By analyzing the circuit, we get

$$V_{B/A}(t) = \frac{1}{2}\alpha_{B/A}(\Delta x)Z_{\text{line}}\left(i_m(t) - i_m\left(t - \frac{L}{v_{\text{beam}}} - \frac{L}{v_{\text{line}}}\right)\right) \quad (\text{A.3})$$

for the voltages at the upstream ports of the transmission lines, with their fourier transforms

$$V_{B/A}(\omega) = \alpha_{B/A}(\Delta x)Z_{\text{line}}I_m(\omega)\exp(j(\pi/2 - \theta))\sin(\theta) \text{ with } \theta = \omega L(1/v_{\text{beam}} + 1/v_{\text{line}})/2. \quad (\text{A.4})$$

To first order, $\alpha_{B/A}(\Delta x) = \alpha_0 \pm \alpha_1 2\Delta x/h$ can be assumed. This gives

$$V_{\Sigma} = \sqrt{\frac{Z_0}{2Z_{\text{line}}}}(V_B + V_A) = 2\alpha_0\sqrt{\frac{Z_0 Z_{\text{line}}}{2}}I_m(\omega)\sin(\theta)\exp(j(\pi/2 - \theta)) \quad (\text{A.5})$$

¹Only for $v_{\text{beam}} = v_{\text{line}}$.

for the voltage at the sum output of the Hybrid and

$$V_{\Delta} = \sqrt{\frac{Z_0}{2Z_{\text{line}}}} (V_B - V_A) = \frac{4\alpha_1 \Delta x}{h} \sqrt{\frac{Z_0 Z_{\text{line}}}{2}} I_m(\omega) \sin(\theta) \exp(j(\pi/2 - \theta)) \quad (\text{A.6})$$

for the voltage of the difference output of the Hybrid. Note that the factor $\sqrt{Z_0/Z_{\text{line}}}$ comes from the fact that the Hybrid may include an impedance transformer. From (2.19) and (2.20) then follow the longitudinal transfer impedance

$$Z_{p\parallel} = \frac{V_{\Sigma}}{I_m(\omega)} = 2\alpha_0 \sqrt{\frac{Z_0 Z_{\text{line}}}{2}} \sin(\theta) \exp(j(\pi/2 - \theta)) \quad (\text{A.7})$$

and the transverse transfer impedance

$$Z_{p\perp} = \frac{V_{\Delta}}{\Delta x I_m(\omega)} = \frac{4\alpha_1}{h} \sqrt{\frac{Z_0 Z_{\text{line}}}{2}} \sin(\theta) \exp(j(\pi/2 - \theta)). \quad (\text{A.8})$$

Note that setting $2\alpha_0 = g_{\parallel}$ and $2\alpha_1 = g_{\perp}$, leads to the same result as in [12]. There are closed form approximations of $g_{\parallel}$ and $g_{\perp}$ depending on the width w of the striplines and the distance h of the striplines to each other

$$g_{\parallel} = \frac{2}{\pi} \arctan \left(\sinh \left(\frac{\pi w}{2h} \right) \right) \quad (\text{A.9})$$

$$g_{\perp} = \tanh \left(\frac{\pi w}{2h} \right). \quad (\text{A.10})$$

These relations are “calculated by conformal mapping used for static fields”[18, p. 6] applied to analytic formulas for the image current on a circular beam pipe. The kicker response is then obtained by using (2.21) and (2.22). The results of the transfer impedance, kick factor and shunt impedance are shown in Tab. 2.1 and Tab. 2.2.

B Classic Multi-Loops

In this Appendix, the transfer impedance, kick factor and shunt impedance of the classic multi-loops in Section 2.2.4 are derived. It is found, that the results are obtained by multiplying the response of a classic stripline device by an array factor.

In the following, the response of a kicker consisting of M stripline devices connected in series by delay lines of arbitrary length (refer to Fig. 2.9) is derived. Results to this problem were published in [17, p. 398] for the case $v_{\text{line}} = v_{\text{beam}} = c$ and $b_i = \lambda/2$ and $a_i = 0$.

[18, pp. 6-7] gives results for the case, when v_{line} and v_{beam} do not equal the speed of light. Another source to be mentioned is [47, pp. 5-6], where the pickup response has been derived for the case $M = 2$, $a_i = 0$ and $b_i = \lambda/2$. The derivation that follows is different from the derivations in [18], [17] and [47] in that it covers arbitrary but equal length of the delay lines.

The circuit diagram in Fig. B.1 is equivalent to the process of the image current of a beam coupling to the structure in Fig. 2.9. The number of elements M can be every natural number and as every element has two ends, there are $N + 1 = 2M$ ends in total.

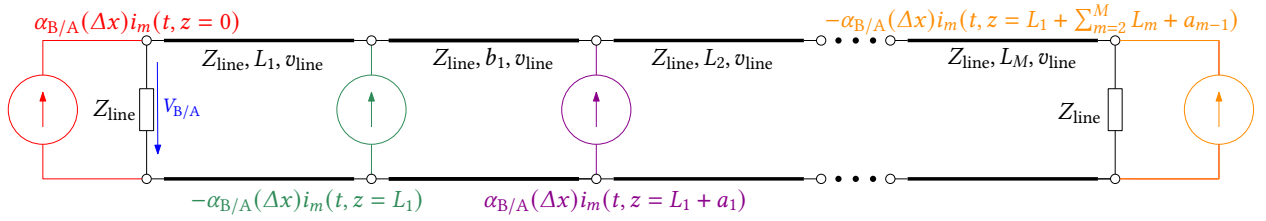


Fig. B.1: Equivalent circuit for the M element multi-loop.

By analysis of the equivalent circuit in (B.1),

$$V_{B/A}(t) = \frac{1}{2} Z_{\text{line}} \sum_{n=0}^N (-1)^n \alpha_{B/A}(\Delta x) i_m(t - t_n) \quad (\text{B.1})$$

follows for the time domain voltages at the upstream ports with their fourier transforms

$$V_{B/A}(\omega) = \frac{1}{2} Z_{\text{line}} \alpha_{B/A}(\Delta x) I_m(\omega) \sum_{n=0}^N \exp(-j\omega t_n - j\pi n). \quad (\text{B.2})$$

By some thinking one can come up with

$$\begin{cases} t_0 = 0 \\ t_n = \frac{n}{2} \left[L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right], & \text{for } n \text{ even.} \\ t_n = \frac{n+1}{2} L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{n-1}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right), & \text{for } n \text{ odd.} \end{cases} \quad (\text{B.3})$$

Using (B.3), leads to the sum in (B.2) being equal to

$$1 + \sum_{n \text{ even}} \exp \left(-j\omega \frac{n}{2} \left[L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right] \right) - \sum_{n \text{ odd}} \exp \left(-j\omega \left[\frac{n+1}{2} L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{n-1}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right) \right] \right) \quad (\text{B.4})$$

$$= \sum_{m=0}^{(N-1)/2} \exp \left(-j\omega m \left[L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right] \right) - \sum_{m=1}^{(N+1)/2} \exp \left(-j\omega \left[mL \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + (m-1) \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right) \right] \right) \quad (\text{B.5})$$

$$= \sum_{m=1}^{(N+1)/2} \exp \left(-j\omega (m-1) \left[L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right] \right) - \exp \left(-j\omega \left[mL \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) + (m-1) \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right) \right] \right). \quad (\text{B.6})$$

Pulling terms that do not depend on m out of the sum and using the fact that $(N+1)/2 = M$ leads to the sum in (B.2) being equal to

$$\left(\exp\left(j\omega\left[L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right]\right) - \exp\left(j\omega\left[\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right]\right) \right) \sum_{m=1}^M \exp\left(-j\omega m\left[L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + \frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right]\right). \quad (\text{B.7})$$

Using

$$\theta = \frac{\omega}{2} L \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) \quad (\text{B.8})$$

$$\zeta = \frac{\omega}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right) \quad (\text{B.9})$$

in (B.7) leads to

$$2j \exp\left(j(\theta + 2\zeta)\right) \sin(\theta) \sum_{m=1}^M \exp\left(-j2m\left[\theta + \zeta\right]\right). \quad (\text{B.10})$$

Using the well known formula for geometric series, (B.10) is equal to

$$2j \exp\left(j(\theta + 2\zeta)\right) \sin(\theta) \exp\left(-j(M+1)(\theta + \zeta)\right) \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)} \quad (\text{B.11})$$

$$= 2j \sin(\theta) \exp\left(j(\zeta - M\theta - M\zeta)\right) \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)} \quad (\text{B.12})$$

$$= 2 \sin(\theta) \exp\left(j\left(\frac{\pi}{2} - \theta\right)\right) \exp\left(-j(M-1)(\theta + \zeta)\right) \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)}. \quad (\text{B.13})$$

Plugging in this result for the sum in (B.2) yields

$$V_{B/A}(\omega) = \underbrace{\alpha_{B/A}(\Delta x) Z_{\text{line}} I(\omega) \exp\left(j\left(\frac{\pi}{2} - \theta\right)\right) \sin(\theta)}_{\text{classic stripline device response}} \underbrace{\exp\left(-j(M-1)(\theta + \zeta)\right) \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)}}_{\text{Array Factor AF}} \quad (\text{B.14})$$

which consists of a factor equal to the classic stripline device response from (A.4) and an array factor

$$\text{AF} = \exp\left(-j\left[(M-1)(\theta + \zeta)\right]\right) \frac{\sin(M[\theta + \zeta])}{\sin(\theta + \zeta)}. \quad (\text{B.15})$$

The pickup and kicker responses of an M element multi-loop are thus equal to the corresponding stripline device response, multiplied by the array factor AF (Tab. B.1). At this point it can be pointed out that the array factor AF has the same form as the array factor of the well known linear antenna array with equidistant elements and constant amplitude of the feed signals.

Tab. B.1: Correspondence of stripline device performance and multi-loop performance. The equations are valid for both transverse and longitudinal components.

Shunt impedance $Z_{\text{shunt, superelec}}$	Kick factor $K_{\text{superelec}}$	Transfer Impedance $Z_{\text{p, superelec}}$
$Z_{\text{shunt, stripline device}} \cdot \text{AF} ^2$	$K_{\text{stripline device}} \cdot \text{AF}$	$Z_{\text{p, stripline device}} \cdot \text{AF}$

C Inverter Multi-Loops

In this Appendix, the response of the inverter double- and inverter triple-loops from Section 4.1 is derived. Specifically, (4.2) and (4.3) are derived. Like for the classic multi-loops, it is found that the response is obtained by multiplying the response of a classic stripline by an array factor.

Inverter multi-loops (refer to Fig. 4.1) are multi-loops with inverter circuits embedded into the lines that connect the striplines. The only difference of inverter multi-loop pickups to classic multi-loop pickups is thus the polarity of the pulses in the time-domain output voltage. The intra-loop array factors for double- and triple-loops are derived in the following. The total multi-loops response is then the response of a single loop times the intra-loop array factor. For both the inverter double- and inverter triple-loops we have

$$\theta = \frac{\omega L}{2} \left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}} \right) \quad (\text{C.1})$$

$$\zeta = \frac{\omega}{2} \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}} \right). \quad (\text{C.2})$$

Inverter double-loop

The intra-loop array factor (4.2) of the inverter double-loop is derived. The voltages at terminals A and B in frequency domain are

$$V_{B/A}(\omega) = \frac{1}{2} Z_{\text{line}} \alpha_{B/A}(\Delta x) I_m(\omega) \underbrace{\sum_{n=0}^3 s_n \cdot \exp(-j\omega t_n)}_{=\Sigma}. \quad (\text{C.3})$$

The coefficients s_n and t_n are shown in Tab. C.1. The sum in (C.3) is equal to

$$\Sigma = 1 - \exp(-j2\theta) - \exp(-j2(\theta + \zeta)) + \exp(-j2(2\theta + \zeta)) \quad (\text{C.4})$$

and can be factorized to the expression

$$\Sigma = \left(1 - \exp(-j2\theta) \right) \cdot \left(1 - \exp(-j2(\theta + \zeta)) \right) \quad (\text{C.5})$$

which leads to

$$\Sigma = \underbrace{2 \sin(\theta) \exp(j(\pi/2 - \theta))}_{\text{single-loop response}} \underbrace{2 \sin(\theta + \zeta) \exp(j(\pi/2 - \theta - \zeta))}_{\text{intra-loop Array Factor}}. \quad (\text{C.6})$$

The intra-loop array factor of an inverter double-loop is thus

$$\text{AF} = 2 \sin(\theta + \zeta) \exp(j(\pi/2 - \theta - \zeta)). \quad (\text{C.7})$$

Inverter triple-loop

The intra-loop array factor (4.3) of the inverter triple-loop is derived. The voltages at terminals A and B in frequency domain are

$$V_{B/A}(\omega) = \frac{1}{2} Z_{\text{line}} \alpha_{B/A}(\Delta x) I_m(\omega) \underbrace{\sum_{n=0}^5 s_n \cdot \exp(-j\omega t_n)}_{=\Sigma}. \quad (\text{C.8})$$

The coefficients s_n and t_n are shown in Tab. C.1. The sum in (C.8) is equal to

$$\Sigma = 1 - \exp(-j2\theta) - \exp(-j2(\theta + \zeta)) + \exp(-j2(2\theta + \zeta)) + \exp(-j4(\theta + \zeta)) - \exp(-j2(3\theta + 2\zeta)) \quad (C.9)$$

and can be factorized to the expression

$$\Sigma = \left(1 - \exp(-j2\theta)\right) \cdot \left(1 - \exp(-j2(\theta + \zeta)) + \exp(-j4(\theta + \zeta))\right) \quad (C.10)$$

$$= \left(1 - \exp(-j2\theta)\right) \cdot \left(\exp(-j2(\theta + \zeta)) - \exp(+j(\pi/3))\right) \cdot \left(\exp(-j2(\theta + \zeta)) - \exp(-j(\pi/3))\right) \quad (C.11)$$

$$(C.12)$$

which leads to

$$\Sigma = \underbrace{2 \sin(\theta) \exp(j(\pi/2 - \theta))}_{\text{single-loop response}} \underbrace{4 \sin(\theta + \zeta + \pi/6) \sin(\theta + \zeta - \pi/6) \exp(j(\pi - 2\theta - 2\zeta))}_{\text{intra-loop Array Factor}}. \quad (C.13)$$

The intra-loop array factor of an inverter triple-loop is thus

$$\text{AF} = 4 \sin(\theta + \zeta + \pi/6) \sin(\theta + \zeta - \pi/6) \exp(j(\pi - 2\theta - 2\zeta)). \quad (C.14)$$

Tab. C.1: Coefficients s_n and t_n for inverter double-loops and inverter triple-loops.

n	s_n	t_n
0	+1	0
1	-1	$L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right)$
2	-1	$L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right)$
3	+1	$2L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + \left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right)$
4	+1	$2L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + 2\left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right)$
5	-1	$3L\left(\frac{1}{v_{\text{beam}}} + \frac{1}{v_{\text{line}}}\right) + 2\left(\frac{a}{v_{\text{beam}}} + \frac{b}{v_{\text{line}}}\right)$

D The AD Stochastic Cooling Kicker Array Factor

In this Appendix, the array factor of the AD stochastic cooling kickers is derived. This Appendix follows the content of Section 2.4, in which the array factor of the AD stochastic cooling pickups was derived.

The derivation of the kicker array factor takes the power splitting network and its in part switchable delays into account. Losses, impedance mismatch and finite bandwidth of the power splitters are not taken into account. This array factor is valid for both kickers¹ and for both the longitudinal and transverse kick factor. The derivation is done for an equivalent pickup and then applied to the corresponding kicker. As mentioned in Section 2.4 this is valid due to reciprocity of pickups and kickers².

The power feeding scheme for one of the two kickers is illustrated in Fig. D.1. Each kicker contains 24 double-loop elements. Per side of one kicker tank, there are 12 feed-ports. Each feed-port delivers power to two double-loop elements with a delay fixed to $\Delta t_{35} = 301$ ps, which is optimal for the first stochastic cooling plateau. Outside of the kicker tank, the following power splitting scheme is used: The 12 feed-ports are grouped into 3 batches. Inside a batch, the delay between two adjacent feed-ports is $2 \cdot \Delta t_{35} = 601$ ps. Between the batches, the delay is switchable between $8 \cdot \Delta t_{35} = 2406$ ps and $8 \cdot \Delta t_{20} = 2570$ ps.

Using the kicker as a pickup, the output voltages $V_{0,1,2}(t)$ of the batches depend on the output voltage $V(t)$ of the double-loops by

$$V_{0,1,2}(t) = \frac{1}{\sqrt{8}} \sum_{k=0}^7 V(t - k(\Delta t_{\text{beam}} - \Delta t_{35})) \quad (\text{D.1})$$

$$V_{0,1,2}(j\omega) = \frac{1}{\sqrt{8}} \sum_{k=0}^7 V(j\omega) \cdot \exp(-j\omega k(\Delta t_{\text{beam}} - \Delta t_{35})) \quad (\text{D.2})$$

The total output voltage depends on the output voltages from the three batches by

$$V_{\text{total}}(t) = \frac{1}{\sqrt{3}} \sum_{m=0}^2 V_{0,1,2}(t - m8(\Delta t_{\text{beam}} - \Delta t_{\text{comp}})) \quad (\text{D.3})$$

$$V_{\text{total}}(j\omega) = \frac{1}{\sqrt{3}} \sum_{m=0}^2 V_{0,1,2}(j\omega) \cdot \exp(-j\omega m8(\Delta t_{\text{beam}} - \Delta t_{\text{comp}})) \quad (\text{D.4})$$

with $\Delta t_{\text{comp}}(\beta = 0.96724) = \Delta t_{35}$ and $\Delta t_{\text{comp}}(\beta = 0.90532) = \Delta t_{20}$. This leads to a total output voltage of

$$V_{\text{total}}(j\omega) = V(j\omega) \underbrace{\frac{1}{\sqrt{24}} \sum_{k=0}^7 \exp(-j\omega k(\Delta t_{\text{Beam}} - \Delta t_{35})) \cdot \sum_{m=0}^2 \exp(-j\omega m8(\Delta t_{\text{Beam}} - \Delta t_{\text{comp}}))}_{\text{AF}} \quad (\text{D.5})$$

The array factor of one stochastic cooling kicker at the two momentum plateaus is thus

$$\text{AF}_{\text{kicker}}(f, \beta_{\text{beam}}) = \begin{cases} \sqrt{24} & \text{for } \beta_{\text{beam}} = 0.96724 \\ \frac{3}{\sqrt{24}} \cdot \frac{\sin(8\pi 21 \text{ ps} f)}{\sin(\pi 21 \text{ ps} f)} \cdot \exp(-j\frac{7}{2}\omega(21 \text{ ps})) & \text{for } \beta_{\text{beam}} = 0.90532 \end{cases} \quad (\text{D.6})$$

At the second stochastic cooling plateau, the array factor has low pass behavior with an absolute value of 4.47 at 1.4 GHz instead of 4.9 which is a reduction of approx. 10 percent in kick factor peak value.

¹Both the KHM and KVM kicker.

²Which is derived from the reciprocity of Maxwells Equations.

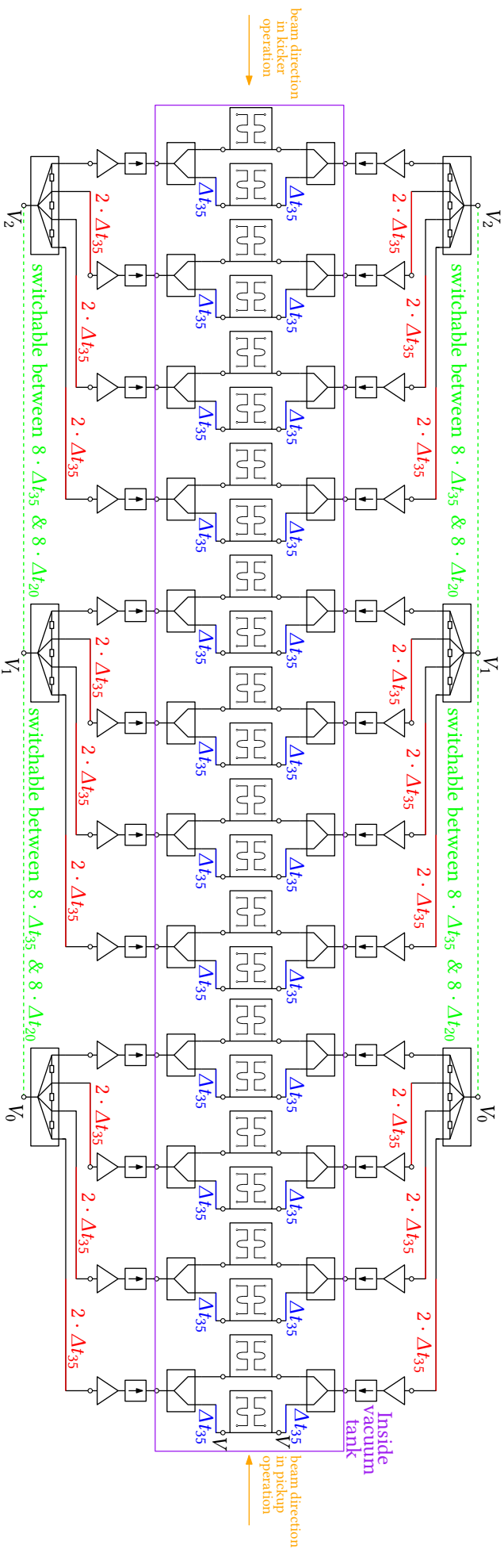


Fig. D.1: Block diagram illustrating the feeding scheme of an AD stochastic cooling kicker. The power splitters inside the vacuum tank are of Wilkinson type without isolation resistors. The power splitters outside the vacuum tank provide isolation between their four output ports. Isolators (terminated circulators) are placed between the power amplifiers and the vacuum tank.

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Acronyms

AD Antiproton Decelerator

AEgIS Antihydrogen Experiment at CERN: Gravity, Interferometry, Spectroscopy [1]

ALPHA Experiment at CERN: The Antihydrogen Laser Physics Apparatus [2]

ASACUSA Experiment at CERN: Atomic Spectroscopy And Collisions Using Slow Antiprotons [3]

BASE Experiment at CERN: The Baryon Antibaryon Symmetry Experiment [4]

CERN European Organization for Nuclear Research

e2ed Electrode to Electrode Distance

ELENA Extremely Low Energy Antiproton ring

EM Electromagnetic

FOM Figure of Merit

GBAR Experiment at CERN: The Gravitational Behaviour of Antimatter at Rest Experiment [5]

GSI GSI Helmholtzzentrum für Schwerionenforschung GmbH, Darmstadt, Germany

GUI Graphical User Interface

KHM Kicker Tank for longitudinal and horizontal kicks

KVM Kicker Tank for longitudinal and vertical kicks

PCB Printed Circuit Board

PEC Perfect Electric Conductor

PS Proton Synchrotron at CERN

SNR Signal-to-Noise Ratio

SPS Super Proton Synchrotron

t-solver Time Domain Solver

TSWS Transverse Standing Wave Device sub-group of slotline devices

TTWS Transverse Travelling Wave Device sub-group of slotline devices

VBA The Visual Basic for Applications programm language