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DEPARTMENT OF RADIO ELECTRONICS

ÚSTAV RADIOELEKTRONIKY

ELECTRONICS FOR SCINTILLATOR DETECTOR READOUT AND COAXIAL RF LINE FOR VITO BEAMLINE AT ISOLDE FACILITY AT CERN

ELEKTRONIKA PRO VYČÍTÁNÍ SCINTILAČNÍCH DETEKTORŮ A KOAXIÁLNÍ RF VEDENÍ PRO
SVAZEK VITO V ZAŘÍZENÍ ISOLDE V CERN

MASTER'S THESIS

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Electronics for Scintillator Detector Readout and Coaxial RF Line for VITO Beamline at ISOLDE Facility at CERN

INSTRUCTION:

This project will be to develop an rf-atom beam interaction region, which will be used to perform rf spectroscopy of short-lived radioactive ion beams delivered at CERN-ISOLDE. Specifically, a vacuum spaced strip-line geometry will be designed and thoroughly simulated. The challenging aspects of this project will be to develop the transition region from strip line to co-axial transmission line geometry while maintaining appropriate impedance matching. Furthermore, the design parameters and constraints relating to the ion-optical properties of the ISOLDE beam and required range of single-mode operation will guide the design choices.

After the initial design phase, the final design of the RF line to be constructed for the experiment will be prepared. The second part of the project will be devoted to designing, building, and validating analog electronics for read-out of a new scintillator detector using Silicon PhotoMultipliers. The new detector will be used in experimental run in 2024 for measurement of magnetic moments of beryllium isotopes.

The electronics design requires low noise, high signal rate capability while ensuring compact design to fit the existing infrastructure. The prototype of the electronics is expected to be ready for testing in spring 2024.

RECOMMENDED LITERATURE:

[1] A. Sen, L. S. Goodman, and W. J. Childs. Collinear Laser and Slow-Ion-Beam Apparatus for High-Precision Laser-RF Double-Resonance Spectroscopy. *Review of Scientific Instruments*, 59(1):74–80, 1988. Available online: <https://pubs.aip.org/aip/rsi/article-abstract/59/1/74/314028>. doi: 10.1063/1.1139969.

[2] M. E. Van Hove and R. E. Silverans. Fast Beam Collinear Laser-RF Double Resonance. *Hyperfine Interactions*, 38(1):773–792, 1987. Available online: <https://link.springer.com/article/10.1007/BF02394874>. doi:10.1007/BF02394874.

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ABSTRACT

VITO (Versatile Ion polarisation Technique Online) beamline experiment is focused on measuring hyperfine structures. RF (Radio Frequency) spectroscopy is used for the measurement. For performing the RF spectroscopy it is necessary to design an RF interaction region for RF spectroscopy of short-lived radioactive atom/ion beams with low energy and verify its parameters and properties by simulation. An important aspect of the RF interaction region is its uniformity. Therefore a vacuum spaced stripline geometry was employed.

VITO beamline also focuses on β -NMR (Nuclear Magnetic Resonance). β -NMR is a method to determine the magnetic moments of short-lived isotopes with very high precision using β particles emitted when the isotopes decay. This master thesis is also focused on the development of electronics for a β particle detector to be used at the β -NMR beamline VITO at ISOLDE (Isotope mass Separator On-Line facility) facility at CERN (Conseil Européen pour la Recherche Nucléaire).

KEYWORDS

Transmission line, stripline, coaxial line, impedance matching, amplifier, RF spectroscopy, β -NMR, SiPM, AD8001, AD8007, VITO beamline, ISOLDE, CERN.

ABSTRAKT

VITO (Versatile Ion polarisation Technique Online) beamline experiment se zabývá měřením hyperjemných struktur. Pro jejich měření se využívá RF (Radio Frequency) spektroskopie. Pro RF spektroskopii je nutné navrhnout a simulacemi ověřit aparaturu pro vytvoření interakční oblasti RF a atomového/iontového svazku umožňující RF spektroskopii krátce žijících radioaktivních atomových/iontových svazků s nízkou energií. Důraz je kladen na uniformní rozložení RF pole. Kvůli potřebě uniformního rozložení RF pole je návrh zaměřen na přenosové vedení typu stripline a kvůli nízké energii krátce žijících radioaktivních svazků je jako dielektrikum zvoleno vakuum.

VITO beamline se rovněž zabývá β -NMR (Nuclear Magnetic Resonance). β -NMR je vysoce přesná metoda určení magnetických momentů krátkodobě žijících izotopů s využitím β částic vyzářených při rozpadu izotopů. Tato diplomová práce se rovněž zabývá návrhem elektroniky pro nový β detektor, který bude využit pro β -NMR beamline VITO v ISOLDE (Isotope mass Separator On-Line facility) zařízení v CERNu (Conseil Européen pour la Recherche Nucléaire).

KLÍČOVÁ SLOVA

Přenosové vedení, stripline, koaxiální vedení, impedanční přizpůsobení, zesilovač, RF spektroskopie, β -NMR, SiPM, AD8001, AD8007, VITO beamline.

ROZŠÍŘENÝ ABSTRAKT

Cílem této práce je navrhnout a odsimulovat počáteční desing RF interakční oblasti pro atomový/iontový svazek a elektroniku pro nový β detektor. RF interakční oblast bude využita při experimentu zabývající se RF spektroskopií a nový β detektor pro β -NMR. Obě tato zařízení jsou navrhována pro VITO beamline v ISOLDE zařízení v CERNu.

Cílem experimentu zabývajícího se RF spektroskopií je měření hyperjemných struktur. Prvním krokem v experimentu je excitace atomů/iontů na vyšší úroveň jemné struktury. K tomu se využívá excitace pomocí laseru. Při opakované excitaci laserem celá populace atomů/iontů je postupně přesunuta na vyšší úroveň jemné struktury. Tento proces se nazývá *Optical pumping*.

Při bližším pohledu lze zjistit, že každá úroveň jemné struktury se dá rozdělit na dvě úrovně hyperjemné struktury. Při aplikaci RF pole o správné frekvenci začne excitovaná populace oscilovat mezi těmito hyperjemnými úrovněmi. Tento proces se nazývá *RF excitation*.

Při opětovné aplikaci laseru dojde k de-excitaci populace atomového/iontového svazku za doprovodu emitovaného fotonu. Tento proces se nazývá *Optical detection* a je použit pro měření hyperjemných struktur.

První část této práce je zaměřena na návrh *RF excitation* sekce a simulace úvodního designu. Požadavky na danou sekci jsou použitelný frekvenční rozsah, uniformní RF pole, minimalizace odrazů a jednoduchá a rychlá výroba. Daný návrh měl být totiž použit při nejbližším beamtimu v listopadu.

Svazek, který bude při experimentu použit, má velice nízkou energii. Z toho důvodu musí být jako dielektrikum použito vakuum. Jakýkoliv jiný materiál, a to včetně vzduchu, by svazek dříve či později zastavil. Uvažovaný frekvenční rozsah pro úvodní design je od 0 Hz do 1 GHz.

Uniformní RF pole je důležité pro správnou excitaci populace atomů/iontů. Cílem je excitovat celou populaci na stejnou hyperjemnou úroveň. Pokud by atomy/ionty viděly v různých částech různé RF pole, došlo by k rozdílné excitaci a atomy/ionty v různých úrovních by byly smíchány dohromady. Z toho důvodu byla zvolena struktura typu stripline.

Jelikož se svazek atomů/iontů pohybuje vůči RF poli, dochází k dopplerovskému posuvu frekvence. Kdyby docházelo k výrazným odrazům v přenosovém médiu, viděl by svazek dvě rozdílné frekvence. Z toho důvodu bylo odsimulováno několik různých geometrií daného přenosového média.

Délka dané RF oblasti je důležitá z důvodu dostatečně dlouhého trvání průletu svazku RF polem. Vhodná délka přenosového média je tedy v místních podmínkách 2 m. Takto dlouhé přenosové médium typu stripline musí být podepřeno, aby se neprohýbalo vlastní vahou. Problematické je podepření zejména vnitřního vodiče.

Pro tento účel byly navrženy teflonové podpory válcového tvaru umístěné mezi užší stranu pásku a stěnu zemní plochy. Dále bude celá stripline umístěna užší stranou vnitřního vodiče směrem dolů, aby se omezilo prohýbání.

Po ověření, že podpory příliš neovlivní vlastnosti přenosového média, bylo doladěno spojení koaxiálního napájení a stripline. Jako nejvhodnější řešení bylo vybráno kolmé připojení k širší straně pásku se středem koaxiálního napájení na hraně vnitřního vodiče stripline.

Kromě tohoto způsobu připojení byly vyzkoušeny i jiné možnosti. Vymodelováno a odsimulováno bylo i připojení k užší straně pásku, které ovšem dosahovalo horších vlastností a připojení přímo k čelu pásku. Připojení k čelu pásku dosahovalo nejlepších vlastností, ale jako takové nemohlo být použito. Koaxiální napájení by překáželo vstupujícímu svazku, který by se nemohl dostat dovnitř struktury. K využití připojení k čelu pásku by bylo potřeba pásek na konci ohnout. To je ovšem výrobně náročnější. Navíc by bylo potřeba vytvořit vhodně zakřivenou zemní plochu, což by vyžadovalo využití CNC (Computer Numerical Control) strojů při výrobě, a to by bylo z časových důvodů problematické. Proto byl i tento způsob připojení zavrhnut pro úvodní návrh.

Dále byl navržen postupný přechod koaxiálního vedení pro vhodný přechod mezi jednotlivými geometriemi a udržení vakua uvnitř přenosového média. Naladěná byla rovněž vzdálenost od konce pásku po vodivou stěnu na konci stripline. Následně byl celý design vymodelován v CAD (Computer Aided Design) softwaru a byly vytvořeny výkresy a podklady pro výrobu.

Bohužel během přípravné fáze experimentu došlo k nehodě a ten byl posunut z listopadu na červen nebo červenec. Bylo tedy rozhodnuto, že úvodní design není potřeba, a nebude tedy vyroben.

Dalším krokem v práci byl návrh a simulace finálního designu. Ten do značné míry vychází z úvodního designu, ale jeho geometrie byla upravena. Požadavky na finální design jsou v podstatě stejné jako na úvodní, ovšem je vyžadován vyšší frekvenční rozsah, a to na rozšíření frekvenčního rozsahu z 1 GHz na alespoň 2.5 GHz.

Pro dosažení většího frekvenčního rozsahu byla změněna a vylepšena v podstatě celá konstrukce. Byla změněna geometrie podpor pro poskytování lepší opory a snížení odrazů signálu. Způsob připojení koaxiálního vedení na stripline bylo rovněž změněno a bylo využito připojení na čelo vnitřního vodiče. Během návrhu bylo vyzkoušeno několik různých možností navázání, což vedlo na dva různé designy, které mohou být použité.

První varianta byla získána ohnutím stripline a následným navázáním koaxiálního vedení na čelo vnitřního vodiče stripline. Výhoda tohoto řešení z mechanického hlediska bylo snazší připojení koaxiálního vedení a možnost využití silnějších vodičů, což je výhodné kvůli velkému přenášenému výkonu přibližně 100 W. Z elektronik-

kého hlediska dosahoval tento design nižších odrazů na nižších kmitočtech, ale byl využitelný jen do frekvence 2.5 GHz. Od této frekvence se začaly objevovat výrazné špičky v S_{11} parametru. Tento frekvenční rozsah je dostatečný, ovšem kvůli možnosti potřeby většího frekvenčního rozsahu v budoucnu byla konstrukce dále upravena.

Druhá varianta opět využívá navázání koaxiálního vedení na vnitřní vodič stripline. Stripline ovšem v tomto případě není zahnutá, ale je rovná. Toto řešení je konstrukčně náročnější kvůli prostorovým omezením limitujícím velikost koaxiálního vedení a samotného přechodu. Toto prostorové omezení je způsobeno otvorem pro vstup svazku, který nesmí být ničím blokován. Toto konstrukční řešení využívá tenčího koaxiálního vedení, což limituje možný použitelný výkon. Ovšem design je využitelný pro požadovaný výkon 100 W. Použité podpory byly rovněž upraveny a místo několika malých podpor byla využita jedna velká po celé délce stripline. S těmito úpravami byl získán vyšší frekvenční rozsah, a to až do 3.5 GHz, ovšem odrazy na nižších kmitočtech jsou pro toto řešení výraznější.

Druhá část práce se zabývá návrhem elektroniky pro nový β detektor. Tento detektor bude využit pro β -NMR. β -NMR je druh NMR, který se využívá pro určení magnetických momentů krátkodobě žijících izotopů s vysokou přesností a citlivostí. Využívá β částic vyzářených při rozpadu radioaktivních izotopů. Cílem těchto detektorů je detekce asymetrií v emisi β částic ve směru 0° a 180° vůči směru magnetického pole od hyperpolarizovaných rozpadajících se izotopů. První plánovaný izotop, který bude měřen v tomto novém detektoru je ^{11}Be .

Celý detektor se skládá z dvou samostatných detektorů - velkého a malého. Každý detektor obsahuje scintilátor. Scintilátor je materiál, který vyzáří foton, když skrz něj projde β částice. Fotony jsou poté detekovány detekčními deskami. Jako detekční element je použit SiPM (Silicon PhotoMultiplier). Jedná se o součástku generující rychlý proudový pulz při absorpci fotonu, což umožňuje měřit i jednotky absorbovaných fotonů s vysokou přesností.

Každý detektor (velký i malý) obsahuje osm detekčních desek a na každé detekční desce je 16 SiPM umístěných na specifických místech. Signály z těchto osmi detekčních desek jsou poté sloučeny pomocí slučovací desky. Signál ze slučovací desky může být volitelně filtrován filtrem typu horní propust, což může být využito při *count-rate* měření, nikoliv však při měření energie. Poté jde signál do DAQ (Data Acquisition System), kde dojde k samotnému změření a uložení výsledků. Součástí návrhu je rovněž deska pro rozvod napájení.

Kvůli potřebě měření energie musí mít navrhnutý systém lineární chování. Měření bude probíhat ve vakuu. Systém tedy musí být kompatibilní s vakuem. To platí hlavně pro velké plastové části trpící jevem *outgassing*. Jedná se o jev, při kterém se uvolňují molekuly plynu uvnitř plastu a znehodnocují kvalitu vakua. Kompatibilita s vakuem byla zajištěna využitím vhodných materiálů a snížení množství použitých

plastů uvnitř vakua. Kvůli umístění systému ve vakuu byla i využita deska pro rozvod napájení, která redukuje počet potřebných feedthrough.

Systém bude rovněž umístěn v magnetickém poli o síle 4.7 T. Je tedy potřeba, aby systém byl nemagnetický, nebo jen velice slabě magnetický. Toho bylo docíleno vhodnou volbou součástek a ověřením pomocí silného magnetu. V budoucnu je uvažováno i ochlazení SiPM kvůli snížení šumu. Desky byly z tohoto důvodu navrhnuty s chladicími pady a s layoutem vhodným pro dobrou distribuci tepla.

Návrh detekčních desek vycházel z designu vytvořeném na Ženevské univerzitě. V návrhu byl využit stejný zesilovač AD8001, jaký byl použit v jejich designu. Obvod byl odsimulován v spice simulátoru s využitím modelu SiPM poskytnutým Ženevskou univerzitou. Model bylo ovšem potřeba upravit, aby splňoval chování našeho konkrétního SiPM.

Po návrhu veškeré elektroniky byl vyroben prototyp detekční desky pro malý detektor. Byly ověřeny magnetické vlastnosti a linearita. Pro ověření linearity bylo využito měření různých zdrojů záření, zejména stroncium a bismut. Byla ověřena linearita měření a navrhnutý systém vyhovuje požadavkům.

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Beamline at ISOLDE Facility at CERN

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Introduction

This master thesis is focused on two different projects for VITO beamline at ISOLDE facility at CERN. The first part is focused on the development of RF interaction region for RF-laser spectroscopy. The second part is focused on the development of electronics for a new β -detector, especially the detection electronics.

The development of the RF-atom/ion beam interaction region is divided into two parts. The initial design that is planned to be used in beamline in November 2023 and the final design that is planned to be used in February 2024. While designing there was an accident and the experiment was postponed.

The RF interaction region will be used to perform RF spectroscopy of short-lived ion/atom beams. The purpose of the experiment is to learn about the distribution of nuclear magnetization in nuclei.

Design is focused on a vacuum spaced stripline geometry with the transition to a coaxial line. Both of the designs, the initial and the final one will be thoroughly simulated and tuned for desired parameters.

For the experiment is important to develop an interaction region with a uniform field. The beam that interacts with the RF field and travels through it should see the same field through the interaction region. Therefore vacuum spaced stripline geometry was suggested as the best option for the experiment. Another key parameter for the interaction region is its length which should be as long as possible considering space constraints. The time that the atom/ion beam spends in the RF field is determined by the length of the interaction region.

Other key parameters are power and frequency range. The power is determined by the magnetic field intensity, that supposed to be around 48 A m^{-1} . Thus the apparatus was designed to 100 W which should produce magnetic field intensity around 80 A m^{-1} . The frequency range for the initial design is up to 1 GHz . For the final design, the frequency range is up to 2.5 GHz .

The most important and challenging aspect of the stripline design from an electrical engineering point of view is the impedance matching of the stripline and the coaxial transmission line. The coaxial transmission line is used for feeding the stripline from an RF generator. The impedance matching should be as good as possible to minimize reflected RF components.

From the mechanical point of view, the most important and challenging aspect is to create the stripline in a vacuum. Also, the design needs to be connected to other parts and devices used in the experiment and the dimensions are also limited by available space. The design needs to be manufacturable in CERN workshops and demountable.

The development of the electronics for new β -detector will be used in β -NMR.

The design is focused especially on detection electronics. In the detector, the β particles are converted to photons with a scintillator and then detected with Silicon PhotoMultiplier. Other parts are the summing card for summing signals from all detection PCBs, power distribution PCB, and filter.

Requirements for the design are linearity, vacuum compatibility, magnetic field compatibility, and low-temperature compatibility. The linearity of the design is important for energy measurements. The electronics will be placed in a vacuum. Vacuum compatibility is especially important for plastic components that suffer the outgassing. The electronics will be placed in a strong magnetic field important for the measurement, thus the design needs to be non-magnetic or very weakly magnetic. Optionally the detection elements will be cooled to lower the noise. The electronics need to be designed for good temperature distribution.

1 VITO beamline experiment

The matter consists of atoms, which have also their own structure. Atoms have two main parts - atomic nucleus and electron shell. The nucleus is made of subatomic particles called protons with positive charge and neutrons without electric charge. The electronic shell is divided into orbitals occupied by electrons. Electrons have a total negative charge equal to that of protons in the nucleus. Both charged subatomic particles are fermions which means that they have their own half-integer intrinsic angular momentum called spin. These spins generate a magnetic field of electrons and nuclei which interact with each other. In case the total spin of the nucleus $I > 0$ and the total spin of the electronic shell is also $J > 0$, the interaction causes the splitting of energy levels of electrons. This phenomenon is called hyperfine splitting or hyperfine structures. [1, 2, 3]

The hyperfine structure can be measured with laser spectroscopy. Laser spectroscopy is a technique used for studying the interaction between light and matter. Laser is used for probing properties of atoms and molecules. The probing is done by examining the way how the particles interact with laser light with a specific wavelength. [4]

The population of atoms/ions is excited with the laser. After some time atoms/ions are de-excited from the upper level to the lower level as can be seen in Fig. 1.1. After several repetitions, all of the population is transferred to a specific hyperfine structure level. This process is called optical pumping. [5, 6, 7]

In Fig. 1.1 there are two fine structure levels ($S_{1/2}$, $P_{1/2}$). When observed with higher resolution these levels are further subdivided into four hyperfine structure levels. The figure is for nuclear spin $I = 3/2$. [2, 3]

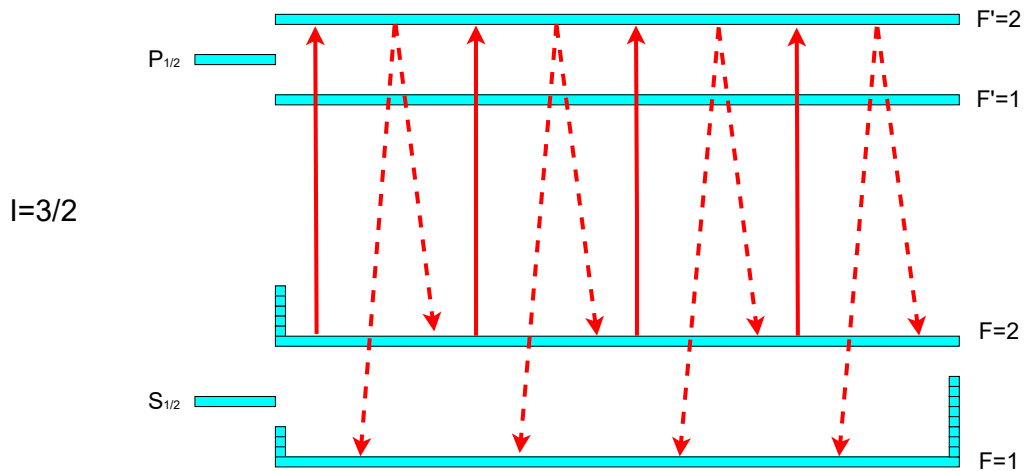


Fig. 1.1: Optical pumping.

After the optical pumping, the RF excitation is applied. When the correct frequency is applied, the population of the hyperfine oscillates between these two levels as can be seen in Fig. 1.2. [7]

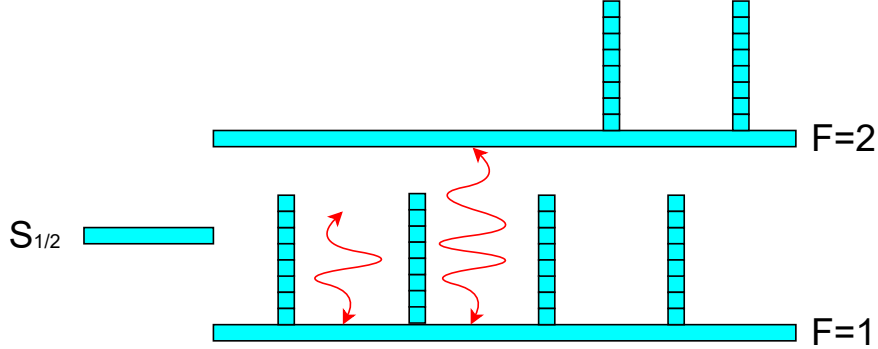


Fig. 1.2: RF excitation.

The transition probability of an atom/ion that interacts during a time t with the external RF field is given by the classical Rabi solution. The mathematical expression for the Rabi function is given as [7]

$$P(\Delta, t) = \frac{\Omega_R^2}{\Omega_R^2 + \Delta^2} \sin^2 \left(\frac{1}{2} t \sqrt{\Omega_R^2 + \Delta^2} \right), \quad (1.1)$$

where $P(\Omega, t)$ is the transition probability, Ω_R is the Rabi frequency which is proportional to the magnitude of the applied field and Δ is the difference of the RF field frequency from the resonance frequency and can be calculated by following equation [7]

$$\Delta = \omega - \omega_0, \quad (1.2)$$

where ω is RF frequency and ω_0 is resonance frequency. [7]

The Rabi frequency Ω_R from equation 1.1 depends on the intensity of RF magnetic field. The visualized atom/ion transition probability can be seen in Fig. 1.3. In the plot 1.3 there are many peaks. For the RF spectroscopy we need to excite all the population. In other words, we need to be on the first peak. This is easily achievable by changing the magnetic field properties. [6, 7]

But the magnetic field needs to be uniform and the atom/ion beam that comes through it needs to see basically the same field everywhere. Otherwise, the decay of the population would differ and the atoms/ions would be in different states and they would be combined and blurred together. And this would strongly affect the measured signal. [6, 7]

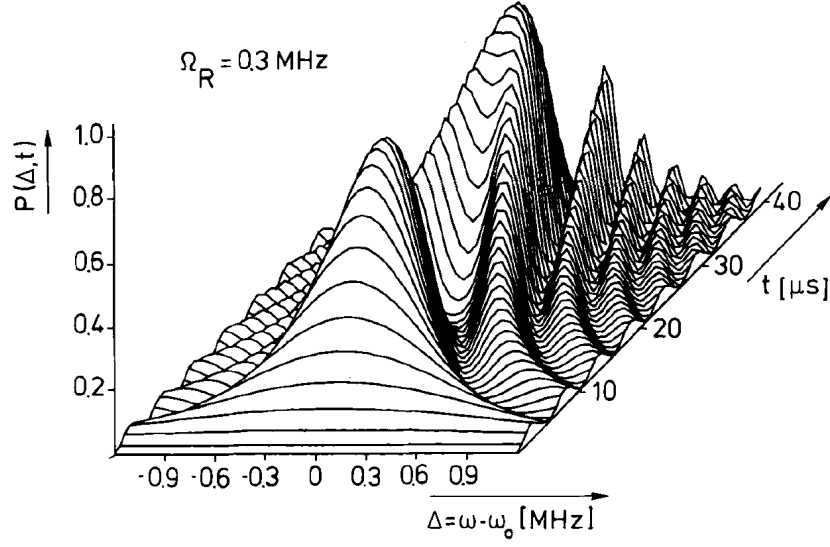


Fig. 1.3: The RF transition probability as a function of the detuning Δ and the interaction time t . [7]

After the RF excitation, the last step is the optical detection which can be seen in Fig. 1.4. Laser is applied again which leads to the excitation of the population to the higher level. Then the population is de-excited again to the lower level and emits photons during this de-excitation. The photons can be then detected. In this way, the laser RF double resonance spectrum can be measured. The width of the peaks is determined by the interaction time with the RF field. [6, 7]

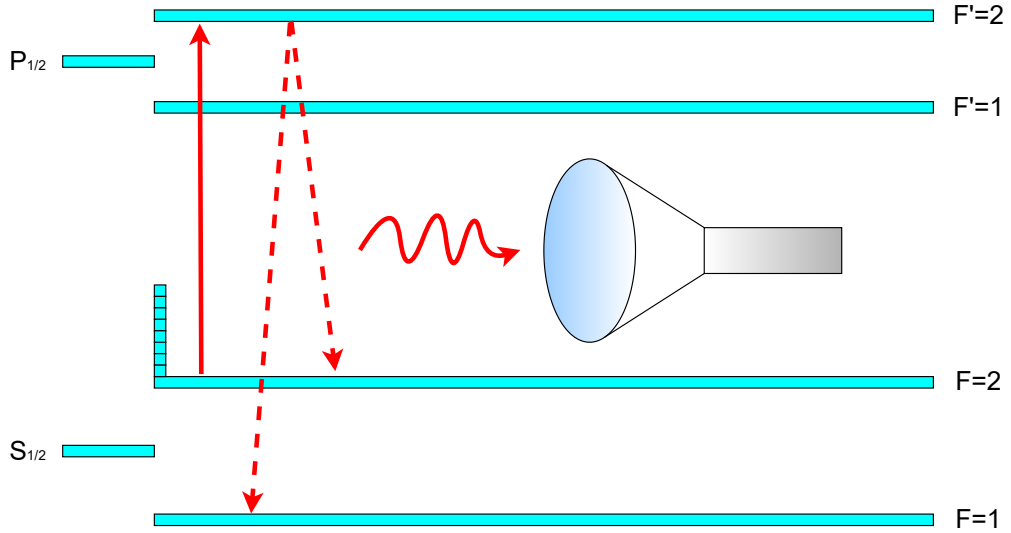


Fig. 1.4: Optical detection.

The whole process summarized into one figure can be seen in Fig. 1.5. At the beginning, the charge exchange cell can be seen. The purpose of the charge exchange cell is to neutralize the ion beam into an atom beam.

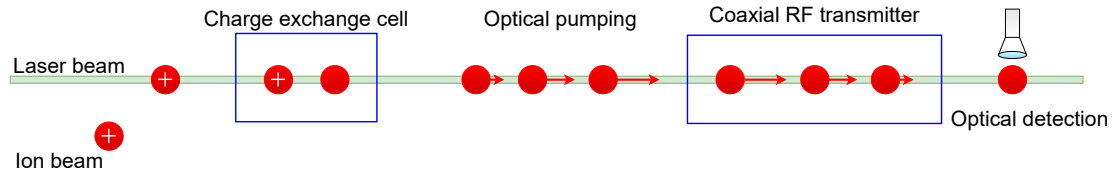


Fig. 1.5: VITO beamline experiment.

2 Signal transmission

By signal transmission, we mean transporting information through a specific medium with some specific quality. The information can be represented in various ways. It can be represented by the amplitude of the signal, by its frequency, by phase change, or by the width of a pulse. It depends on the technology we are using and on the specific application. The signal itself can be analog or digital and is usually represented by electrical voltage. But it can be other physical quantity, for example, electrical current.

The medium itself has some parameters too. The medium can be wired or made from a conductor, wireless, or fiber-optic. This chapter is focused on transmission media made from conductors, on their types and description.

2.1 Basic types of transmission media made from conductors

We can divide transmission media made from conductors into two basic categories - two wire transmission lines and waveguides. Each of them has different parameters and usage and can be further divided. [8]

Some examples: [8, 9]

- two wire transmission lines,
 - open,
 - coaxial,
 - MIC (Microwave Integrated Circuit),
 - * microstrip,
 - * stripline,
 - * coplanar strips,
 - * coplanar waveguide,
- waveguides,
 - rectangular waveguide,
 - circular waveguide,
 - single ridged waveguide,
 - double ridged waveguide,
 - SIW (Substrate Integrated Waveguide).

2.2 Coaxial line

The coaxial line consists of two conductors - inner conductor and outer shielding. The inner conductor is inserted into the outer shielding. Between these two conductors, there is an inner dielectric insulation. The materials commonly used for dielectric insulation are polyethylene, foam polyethylene, and teflon. Air, vacuum, or a combination of different materials can be used. The last layer of coaxial cable is the outer jacket. The outer jacket has no electrical function. It protects the cable against for example moisture, dirt, and abrasion. The construction of coaxial cable can be seen in the following Fig. 2.1. [10]

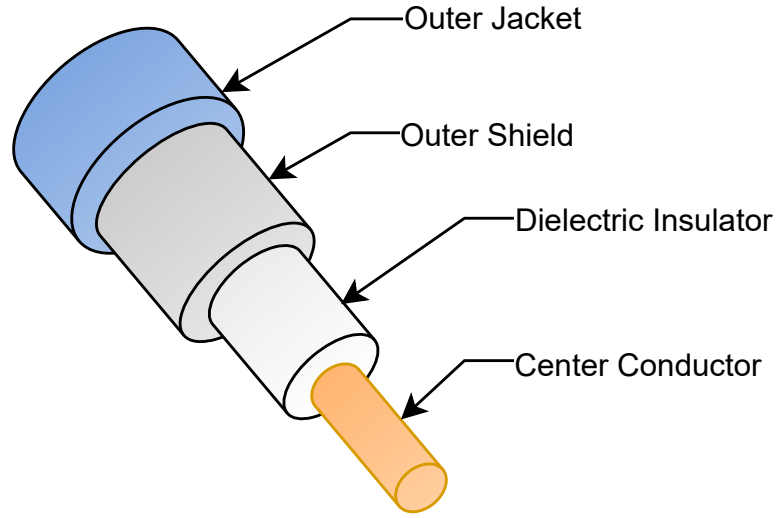


Fig. 2.1: Construction of coaxial cable.

The dominant mode of the coaxial line is TEM (Transverse ElectroMagnetic). The cutoff frequency for TEM wave is 0 Hz thus DC (Direct Current) signal can propagate through the coaxial line. The electric and magnetic field distribution can be seen in Fig. 2.2. [8, 9]

The characteristic impedance of the coaxial line is determined by the dielectric relative permittivity of the dielectric insulation, the diameter of the inner conductor, and the diameter of the inner part of the outer conductor. The coaxial line impedance can be calculated by the following equation [8, 9]

$$Z_{CL} = \frac{60}{\epsilon_r} \cdot \ln \frac{R_0}{r_0}, \quad (2.1)$$

where Z_{CL} is coaxial line impedance, ϵ_r is relative permittivity of the dielectric material, R_0 is inner radius of outer conductor (shielding) and r_0 is inner conductor radius. [8]

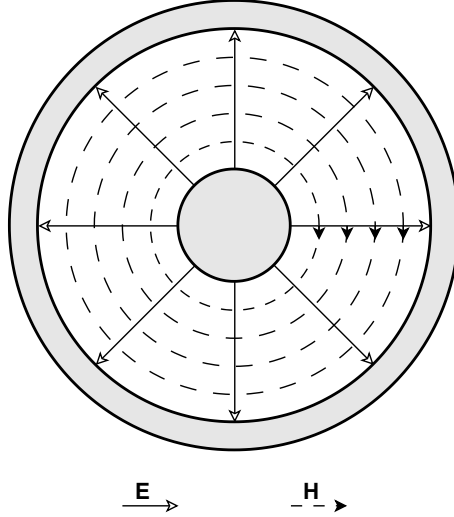


Fig. 2.2: Dominant TEM mode field distribution in coaxial line.

The maximum real power propagated by TEM wave is determined by the inner conductor diameter, the diameter of the inner part of the outer conductor, and the maximum electric field intensity on the surface of the inner conductor and it is determined by following equation [8, 9]

$$P_0^{TEM} = \pi r_0^2 \cdot \sqrt{\frac{\mu}{\varepsilon}} \cdot E_{max}^2 \cdot \ln \frac{R_0}{r_0}, \quad (2.2)$$

where P_0^{TEM} is maximal power, E_{max} is maximal electric field intensity between the two conductors, ε is permittivity and μ is permeability. [8]

The maximum real power cannot exceed the dielectric breakout threshold of the dielectric insulator. The maximum of real power is reached when the ratio of the diameter of the inner part of the outer conductor to the diameter of the inner conductor $R_0/r_0 = 1.65$. Thus if we want to transmit maximum power the multiplication of characteristic impedance and the square root of the insulator's relative permittivity should be $Z_{CL} \cdot \sqrt{\varepsilon_r} = 30 \Omega$. [8, 9]

Loss of coaxial line is determined by the diameter of the inner conductor, the diameter of the inner part of the outer conductor, and surface resistivity and can be calculated by the following equation [8, 9]

$$\alpha_{TEM} = \frac{R_S}{\sqrt{\frac{\mu}{\varepsilon}}} \cdot \frac{1}{2R_S} \cdot \frac{1 + \frac{R_0}{r_0}}{\ln \frac{R_0}{r_0}}, \quad (2.3)$$

where α_{TEM} is the loss and R_S is surface resistivity which can be calculated by following equation [8]

$$R_S = \sqrt{\frac{\omega \mu_v}{2\sigma_v}}, \quad (2.4)$$

where ω is frequency (in radians per second) and σ_v is conductance of the conducting material. [8]

The minimum of loss is reached when the ratio of the diameter of the inner part of the outer conductor to the diameter of the inner conductor $R_0/r_0 = 3.6$. Thus if we want minimal loss the multiplication of characteristic impedance and the square root of the insulator's dielectric relative permittivity should be $Z_{CL} \cdot \sqrt{\epsilon_r} = 77 \Omega$. For commonly used dielectric insulator polyethylene the relative dielectric relative permittivity is $\epsilon_r \approx 2.4$. Thus the ideal characteristic impedance is $Z_{CL} \approx 50 \Omega$. [8, 9]

2.3 Microstrip line and stripline

Both microstrip line and stripline are HMIC (Hybrid Microwave Integrated Circuit). They consist of a conductor, ground plane, and dielectric substrate that separates the conductor from the ground plane. Commonly used materials for the substrate are for example FR4, Rogers, or Arlon. [8]

The difference between a microstrip line and a stripline is that a stripline conductor is embedded into the substrate and it has a ground plane from both sides - top and bottom. Contrary to stripline microstrip line conductor is on the substrate. The difference can be seen in Fig. 2.3. In the case of stripline the conductor is surrounded by the substrate and in the case of microstrip line the substrate is only under the conductor. Above the conductor is usually air. [8]

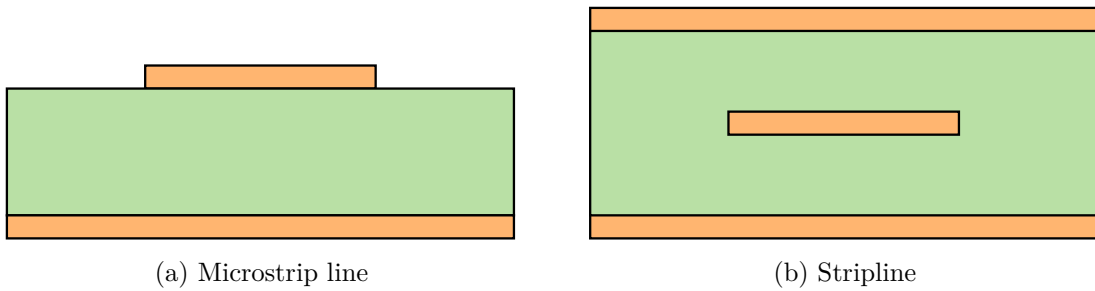


Fig. 2.3: Microstrip line and stripline.

A few different formulas can be used to design microstrip lines. None of these formulas are exactly precise and they have some limitations. Choosing a different formula can lead to slightly different results. Using an online internet calculator specialized for designing microstrip lines is also an option. For getting the best

possible results the design needs to be fine-tuned in some simulator such as Ansys HFSS or CST Studio. [8, 9]

The electric and magnetic field distribution in the stripline and microstrip line is a bit similar and can be seen in Fig. 2.4. The stripline magnetic field has a stadium shape. Its electric field looks like lines from the inner conductor to the ground plane. The lines are straight in the center and curvy on the edges. In the case of a stripline, the field distribution is horizontally and vertically symmetrical. In the case of a microstrip line, the fields are similar to stripline but the fields are a little bit deformed and not horizontally symmetrical. It is caused by the nonsymmetrical structure of the microstrip line and the missing substrate and ground plane above the conductor. [8, 9]

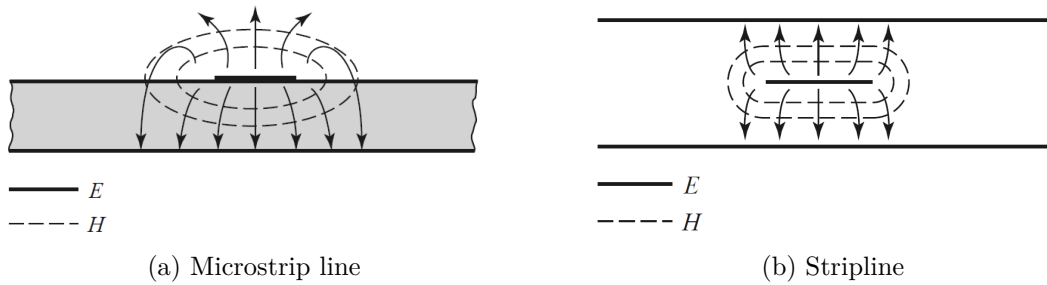


Fig. 2.4: Microstrip line and stripline field distribution. [9]

2.4 Waveguides

A waveguide consists of a hollow conductor of various shapes, usually rectangular, filled with dielectric. Other popular shapes are circular, single ridged, or double ridged shape as can be seen in Fig. 2.5. As the dielectric air is usually used. [8]

Waveguides are a type of transmission line that can propagate only waves with higher frequencies than its cutoff frequency. Lower frequencies are significantly attenuated. Higher frequencies can propagate with almost no attenuation. The transition between passband and stopband is very steep. Thus cutoff frequency is the lowest frequency that can propagate through the waveguide. [8, 9]

The shape of a waveguide together with its size determines the cutoff frequency and dominant mode of the waveguide. In general infinite number of TE (Transverse Electric) and TM (Transverse Magnetic) modes can propagate through the waveguide with its dominant mode on the cutoff frequency. [8, 9]

The size of a waveguide is similar to the wavelength of wave propagating through the waveguide. For smaller frequencies when the wavelength is big the waveguide

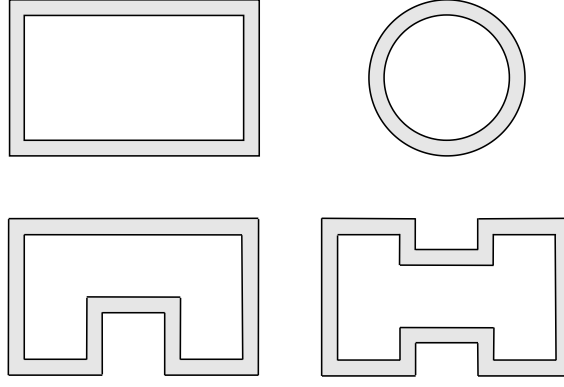


Fig. 2.5: Different waveguides shapes.

would be big as well. Thus it is not very useful for lower frequencies. The distribution of electric and magnetic field in waveguide depends on its shape and mode of the propagating waves. In the following Fig. 2.6 can be seen an example of the field distribution in a rectangular waveguide with wave mode TE_{10} . [8, 9]

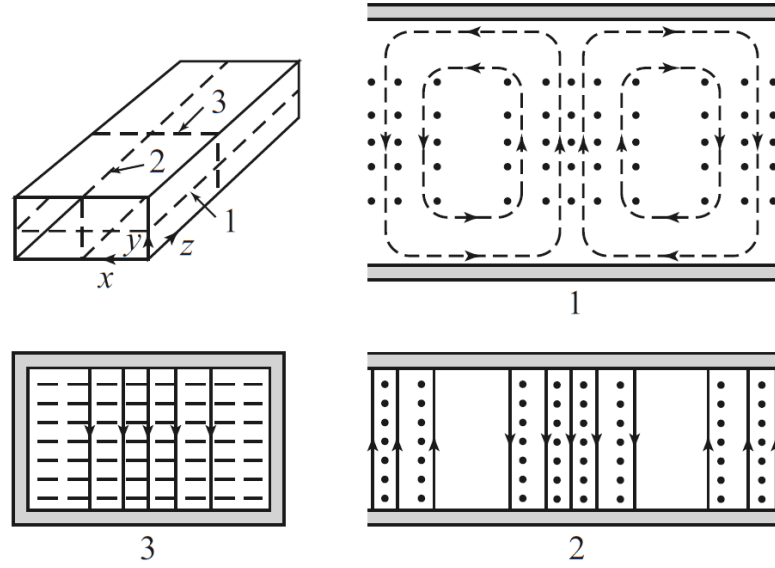


Fig. 2.6: An example of waveguide field distribution. [9]

2.4.1 Substrate Integrated Waveguide

SIW (Substrate Integrated Waveguide) is special type of waveguide. It is basically a combination of waveguide and microstrip line and it combines some of their features. SIW is composed of two conducting planes. Between those planes, there is a dielectric substrate. Both of the planes are connected by two lines of vias close to the

edges of the conducting planes and dielectric substrate which forms a shape similar to rectangular waveguide. An illustration of SIW can be seen in the following Fig. 2.7. The difference between SIW and the classical waveguide is that SIW supports only TE wave modes. This is caused by the discontinuity of the side walls made by the vias. [11]

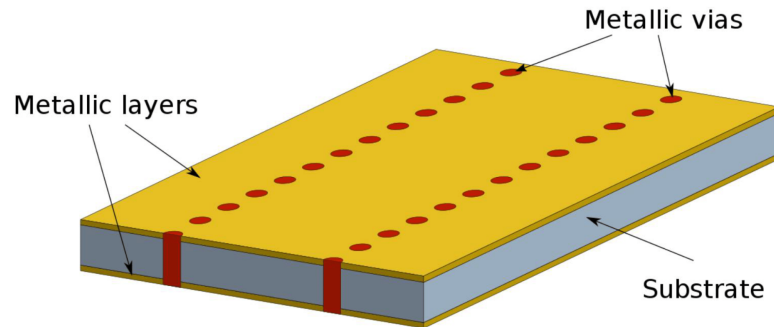


Fig. 2.7: SIW. [11]

This combination of microstrip line and waveguide brings up many advantages. It is small and lightweight, easy to manufacture with less resources. It can be simply integrated and connected to other structures such as microstrip lines. [11]

Also, it is possible to use SIW with higher frequencies than microstrip lines and with greater power, less losses, and smaller attenuation. But it suffers greater losses and attenuation than classic waveguide and cannot handle such big powers. [11]

2.5 Comparison

If we compare a coaxial line to a waveguide we can say that waveguide can have lower losses and can transmit much higher power in specific cases. On the other hand impedance matching of waveguide is much more frequency dependent, coaxial line has wider bandwidth, and waveguide size is strongly dependent on a wavelength. Thus on higher frequencies waveguide is often used and on lower frequencies is preferred coaxial line. [8]

If we compare microstrip line and stripline to coaxial cable it has higher losses and can transmit lower power. However, it is smaller with less weight with easy fabrication and integration. Microstrip line and stripline has higher bandwidth than waveguides. But cannot be used with such high frequencies as waveguides because of dispersion. At higher frequencies, the parameters of the microstrip line and stripline begin to be much more frequency-dependent. [8, 12]

2.6 Impedance matching

Loads in general can have complex impedance Z that consists of real part R that is called resistance and imaginary part X that is called reactance. The impedance is then $Z = R + jX$. If a load with some impedance is connected to a generator with a different impedance, reflections will occur. This reflection can be calculated by the following equation [13]

$$\Gamma_L = \frac{Z_L - Z_S}{Z_L + Z_S}, \quad (2.5)$$

where Γ_L is reflection coefficient of the load, Z_L is load impedance and Z_S is source or generator impedance. Similarly, the reflection coefficient of the source Γ_S can be calculated. The visualization can be seen in Fig. 2.8. This means that a part of the signal will get to the load and part of the signal is reflected back to the generator. [13]

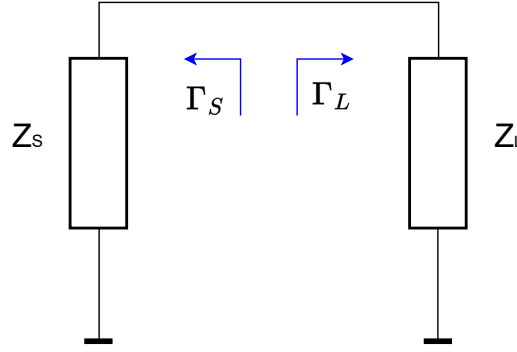


Fig. 2.8: Reflection coefficient.

Impedance matching occurs when the impedance of the load is the same as the impedance of the source. In that case, the reflection coefficient is zero and the whole signal gets to the load and nothing returns back. This is the desired state. In this case, there is only one wave traveling through the transmission line and no reflected wave. The power loss of the transmission of the electromagnetic wave is minimized, the voltage and current across the transmission line is the lowest with the specific power and it may improve SNR (Signal to Noise Ratio). [9, 13]

The impedance matching cannot be perfect and there always will be reflected wave. The goal is to keep the reflected wave as low as possible. By the interaction of the wave going from a source to the load and the reflected wave standing wave appears on the transmission line. A standing wave is a wave whose minimums and maximums of amplitude are constant in space. The bigger the reflected wave is, the higher the amplitude of the standing wave. [13]

The quality of impedance matching can be described by the reflection coefficient or SWR (Standing Wave Ratio). The SWR can be calculated by following equation [13]

$$SWR = \frac{1 + |\Gamma_L|}{1 - |\Gamma_L|}. \quad (2.6)$$

Ideally, the reflection coefficient is equal to 0, and thus the SWR is equal to 1. In reality, this is impossible and the goal is to keep these parameters as low as possible. [13]

The impedance matching can be achieved in many ways. One option is to use lumped elements. They are capacitors, inductors, or resistors connected to matching network. The simplest matching network is the L-section. Other options are T-section or Π -section. [9]

Another option is to use segments of transmission lines and transformers or series of shunt stubs. In our case, the impedance matching is done by tuning the dimensions of the stripline to be $50\ \Omega$ and selecting the appropriate way of connecting the coaxial line that is also $50\ \Omega$. [9]

3 Amplifiers

The amplifier is a component that increases the power, voltage, or current of a signal. The ideal amplifier has infinite input resistance. It draws no current from the source, ensuring that the input signal is not affected. The output resistance of the ideal amplifier is zero. It can deliver its maximum output voltage or current to the load without being affected by the load itself. [14, 15]

There are many important parameters for amplifiers. Some of the main are: [14, 16]

- gain,
- frequency response and bandwidth,
- input and output impedance,
- noise,
- power handling,
- efficiency.

3.1 Basic amplifier types

There are 4 basic types of amplifiers: [14, 17]

- voltage amplifier,
- current amplifier,
- transconductance amplifier,
- transresistance amplifier.

As the name implies voltage amplifier amplifies input voltage to output voltage. The current amplifier does the same but with current. Transconductance amplifiers convert input current to output voltage and transresistance amplifiers work the other way around (convert input voltage to output current). [14, 17, 18]

Except for this, we can also divide amplifiers by their operation, application, or characteristics. Some of the most common types are: [14, 17, 18]

- operational amplifier,
- power amplifier,
- instrument amplifier,
- transistor amplifier,
- vacuum tubes amplifier,
- RF amplifier,
- audio amplifier,
- high-speed amplifier,
- low noise amplifier.

3.2 Summing amplifier

The summing amplifier is a type of electronic circuit. It contains a resistive network connected to the input of an amplifier with a feedback resistor. We can use inverting or non-inverting amplifier input leading to inverting or non-inverting summing amplifier. An example of an inverting summing amplifier can be seen in figure 3.1. [14, 19]

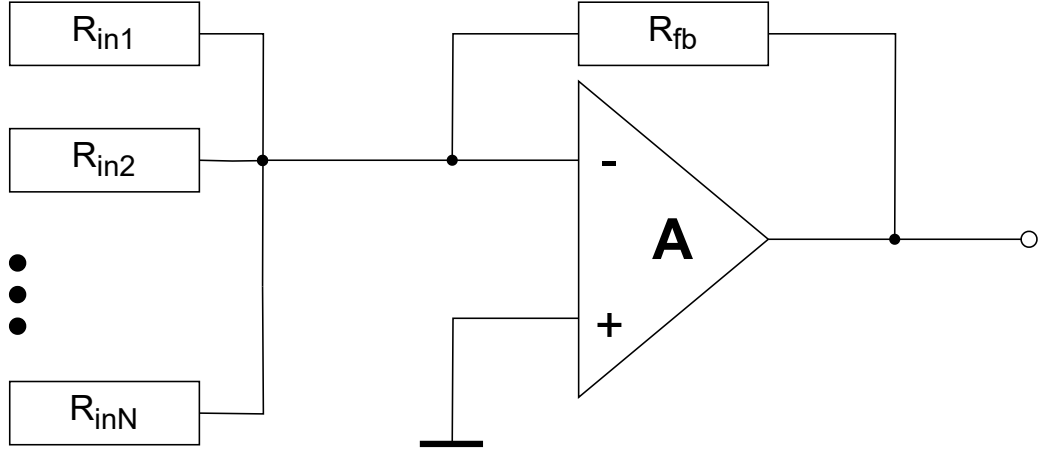


Fig. 3.1: Inverting summing amplifier.

A simple inverting voltage summing amplifier combines voltages connected to the inverting input of the amplifier via a resistor network. The amplification of each input voltage is given by the ratio of feedback resistor and input resistor. The output voltage can be determined by the following equation [14, 19]

$$V_{out} = -R_{fb} \left(\sum_{i=1}^N \frac{V_i}{R_i} \right), \quad (3.1)$$

where V_{out} is output voltage, R_{fb} is feedback resistor, V_i represents input voltages, R_i represents input resistors and N is the total number of input voltages. [14, 19]

4 Initial design

The first part of the thesis is focused on designing a coaxial RF line for VITO beamline. For the development of the RF-atom/ion beam interaction region, we have chosen a vacuum spaced stripline geometry as the best solution. It was selected especially due to its uniform field distribution so the RF-atom/ion beam sees the same RF field in its whole cross section. The vacuum is necessary due to the low energy of the beam. The energy of the beam is usually between 30 keV and 50 keV. The mean free path of particles with such energy in the air would probably be below 1 mm.

4.1 Stripline design

The first step was to design the stripline itself. To get the desired dimensions calculation can be done by hand or a calculator (online or software calculator) can be used. But each source has a little bit different equations and leading to a slightly different results. In this subchapter, some of these equations and different calculators will be compared.

In the following equations Z_0 is the desired impedance (that is $50\ \Omega$), w is the width of the center conductor, t is the thickness of the center conductor, h is the height of a substrate, ϵ_r is relative permittivity, w_{ef} is effective width of the center conductor, H is the distance between center conductor and ground plane and C_f is fringing capacitance at the edges of the strip. The dimensions and fringing capacitance illustration can be seen in Fig. 4.1 [20]

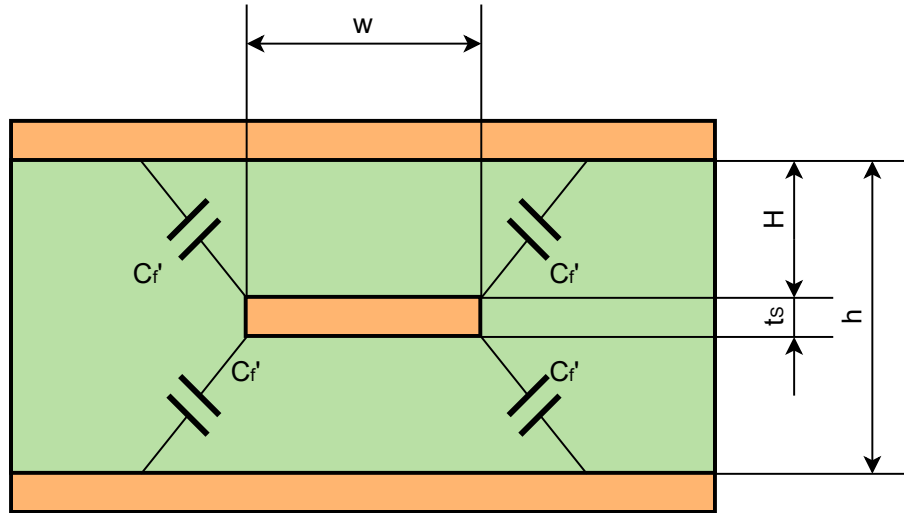


Fig. 4.1: Stripline dimensions and fringing capacitance.

Microwave Engineering [9]

The first step is to calculate the effective width of the center conductor by the following equation [9]

$$w_{ef} = w - \begin{cases} 0 & \text{for } \frac{w}{h} > 0.35 \\ h \cdot \left(0.35 - \frac{w}{h}\right)^2 & \text{for } \frac{w}{h} < 0.35 \end{cases}. \quad (4.1)$$

The equation for this depends on the ratio of the width of the center conductor and the substrate's height. With effective width, the impedance can be then calculated by the following equation [9]

$$Z_0 = \frac{30\pi}{\sqrt{\varepsilon_r}} \cdot \frac{h}{w_{ef} + 0.411h}. \quad (4.2)$$

These formulas assume a strip with zero thickness and are quoted as being accurate to about 1 % of the exact results [9]. The designed stripline center conductor will be quite thick thus some deviations can be expected.

Analog Devices [21]

This equation is taken from Analog Devices article and the impedance is directly calculated from the stripline dimensions by the following equation [21]

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left(\frac{1.9h}{0.8w + t_s} \right). \quad (4.3)$$

IEEE article Stripline Design and LibreTexts Engineering [22, 20]

These equations can be found in two sources. One is IEEE (Institute of Electrical and Electronics Engineers) article about stripline design and one is on LibreTexts Engineering webpage. First, the effective width needs to be calculated by the following equation [22, 20]

$$w_{ef} = w - \begin{cases} 0 & \text{for } \frac{w}{h} \geq 0.35 \\ h \cdot \frac{\left(0.35 - \frac{w}{h}\right)^2}{1 + \frac{12t_s}{h}} & \text{for } \frac{w}{h} < 0.35 \end{cases}. \quad (4.4)$$

The next step is to calculate the fringing capacitance by the following equation [20, 22]

$$C_f = \frac{2}{\pi} \cdot \ln \left(\frac{1}{1 - \frac{t_s}{h}} + 1 \right) - \frac{t_s}{h\pi} \cdot \ln \left[\frac{1}{\left(1 - \frac{t_s}{h}\right)^2} - 1 \right]. \quad (4.5)$$

Now the impedance can be calculated by the following equation [20, 22]

$$Z_0 = \frac{30\pi}{\sqrt{\varepsilon_r}} \cdot \frac{1 - \frac{t_S}{h}}{\frac{w_{ef}}{h} + C_f}. \quad (4.6)$$

Microwaves101 [23]

This equation is taken from Microwaves101 article and the impedance is directly calculated from the stripline dimensions by the following equation [23]

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left[\frac{4h}{0.67w\pi \left(0.8 + \frac{t_S}{w} \right)} \right]. \quad (4.7)$$

Altium Designer [24]

The next equations are from Altium Designer article. The equation is again separated for wide and narrow striplines. First, the effective width needs to be calculated by the following equation [24]

$$w_{ef} = w - \begin{cases} 0 & \text{for } \frac{w}{h} > 0.35 \\ w + \frac{t_S}{\pi} \cdot \ln \left[\frac{\varepsilon_r}{\sqrt{\left(\frac{t_S}{4H + t_S} \right)^2 + \left(\frac{t_S\pi}{4w + 4.4t_S} \right)^m}} \right] & \text{for } \frac{w}{h} < 0.35 \end{cases}, \quad (4.8)$$

where H can be calculated in by following equation

$$H = \frac{h - t_S}{2}, \quad (4.9)$$

and m can be calculated by following equation [24]

$$m = \frac{6H}{3H + t_S}. \quad (4.10)$$

The next step is to calculate fringing capacitance by the following equation [24]

$$C_f = \frac{2h}{h - t_S} \cdot \left[\ln \left(\frac{h}{h - t_S} + 1 \right) - \frac{t_S}{h - t_S} \cdot \ln \left(\frac{1}{1 - \left(\frac{t_S}{h} \right)^2} - 1 \right) \right]. \quad (4.11)$$

This step is necessary only for the wide stripline. Now the final step is to calculate the impedance with equation [24]

$$Z_0 = \begin{cases} \frac{94.15}{\sqrt{\varepsilon_r} \left(\frac{w}{h - t_s} + \frac{C_f}{\pi} \right)} & \text{for } \frac{w}{h} > 0.35 \\ \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left[1 + \frac{8h}{w_{ef}\pi} \left(\frac{16h}{w_{ef}\pi} + \sqrt{\left(\frac{16h}{w_{ef}\pi} \right)^2 + 6.27} \right) \right] & \text{for } \frac{w}{h} < 0.35 \end{cases} \quad (4.12)$$

PE Pasternack [25]

Here is another equation that allows us to calculate the impedance of the stripline directly from its diameters. The equation is taken from an article on PE Pasternack webpage and it is [25]

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left[\frac{1.9(2H + t_s)}{0.8w + t_s} \right], \quad (4.13)$$

where H can be calculated with use of equation 4.9.

EEWeb [26]

The last considered approach to calculate impedance is taken from an article on EEWeb webpage. This approach is again divided into narrow and wide stripline. The impedance can be then calculated by equation [26]

$$Z_0 = \begin{cases} \frac{94.15}{\frac{w}{h} + \frac{\theta}{\pi}} & \text{for } \frac{w}{h} \geq 0.35 \\ \frac{60}{\sqrt{\varepsilon_r}} \cdot \ln \left(\frac{4h}{D\pi} \right) & \text{for } \frac{w}{h} < 0.35 \end{cases}, \quad (4.14)$$

where D is coefficient for simplification of the impedance calculation equation 4.14 and can be calculated by following equation [26]

$$D = \frac{w}{2} \left[1 + \frac{t_s}{w\pi} \left(1 + \ln \left[\frac{4w\pi}{t} \right] \right) + 0.551 \cdot \left(\frac{t_s}{w} \right)^2 \right] \quad (4.15)$$

and θ is coefficient for simplification of the impedance calculation equation 4.14 and can be calculated by following equation [26]

$$\theta = \frac{2h}{h-1} \cdot \ln \left(\frac{2h-1}{h-1} \right) - \frac{t_s}{h-1} \cdot \ln \left[\frac{2ht_s - t_s^2}{(h-t_s)^2} \right] \quad (4.16)$$

Comparison

For the comparison of the results of each equation, we have created a Matlab script so the results can then be compared together with results from different online

calculators or other software calculators. Each calculation has the same input parameters that can be seen in Table 4.1. These parameters are already fine-tuned in CST Studio.

Table 4.1: Input parameters

Magnitude	Value
w	34.8 mm
h	32 mm
t	4 mm
l	2000 mm
ϵ_r	1
Z_0	50.0685 Ω

The results of each calculation can be seen in Table 4.2. The result of CST Studio simulation is considered the most precise and exact result. The deviation in the table 4.2 is the difference between the specific calculation result and CST Studio simulation result.

Table 4.2: Comparison of results of different stripline equations and calculators

Type	Name	Result [Ω]	Deviation [Ω]	Source
hand calculation	Microwave Engineering	61.6603	11.5918	[9]
	Analog Devices	38.8120	-11.2565	[21]
	IEEE, LibreTexts	50.9119	0.8434	[20, 22]
	Microwaves101	38.8233	-11.2452	[23]
	Altium Designer	49.0262	-1.0423	[24]
	PE Pasternack	84.0383	33.9698	[25]
	EEWeb	75.8047	25.7362	[26]
online calculator	Microwaves101	50.66	0.5915	[27]
	Altium Designer	50.9	0.8315	[28]
	PE Pasternack	84.04	33.9715	[25]
	EEWeb	87.65	37.5815	[26]
	All About Circuits	84.6	34.5315	[29]
software calculator	KiCAD	50.8767	0.8082	-
	AppCAD	51.2	1.1315	-
	CST Studio	50.86	0.7915	-

From the deviation of the results in Table 4.2 the results can be divided into three categories. In the first category, the results are close to the required value of

50 Ω with only small differences. To this category belongs all software calculators as expected and only some online calculators or equations. All of these calculators are usable in the stripline design.

The second category has its results further from the assumed right value. The absolute difference is bigger than 10 Ω . This calculators may work in different cases but in this case, they are far from the required value. These equations or calculators might be overly simplified or they might work only for thin striplines.

The third category of results seems to be completely wrong. The results differ from 25 Ω up to almost 40 Ω . These calculators or equations will probably not work in any case.

After the calculation, we designed the stripline in CST Studio and fine-tuned it. The diameters of the stripline can be seen in Table 4.1 together with its impedance. The height of the substrate h and the thickness of the inner conductor t_S were selected according to available material in CERN store¹. The outer conductor of the stripline is made from an aluminum rectangular tube and the inner conductor is made from an aluminium rectangular rod. The diameters of the stripline can be seen in Fig. 4.2. The length l of the stripline itself is 2 m. This length is chosen to create long enough RF-atom/ion beam interaction region.

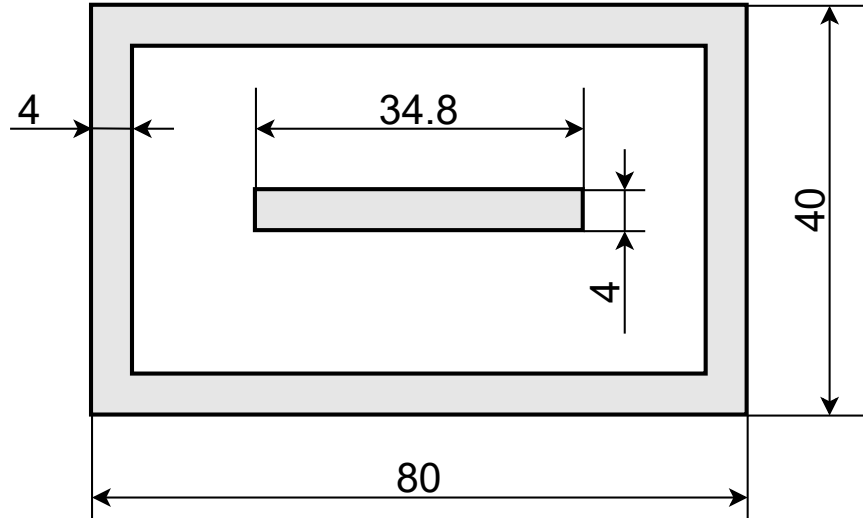


Fig. 4.2: Stripline diameters.

For the initial design the considered frequency range is below 1 GHz. In this band, the stripline works fine as can be seen in Fig. 4.3. The S_{11} parameter is below -130 dB in considered band and the S_{21} parameter is basically 0 dB. For

¹CERN store carries various materials and components in stock, this is relevant due to the rather short time-scale of the project as external delivery times can be significant

the dielectric we used vacuum and for the conductive parts, we used PEC (Perfect Electronic Conductor). Thus the results can be expected slightly worse in reality. The next step is to create a transition between the stripline and a coaxial line that will serve as a feed.

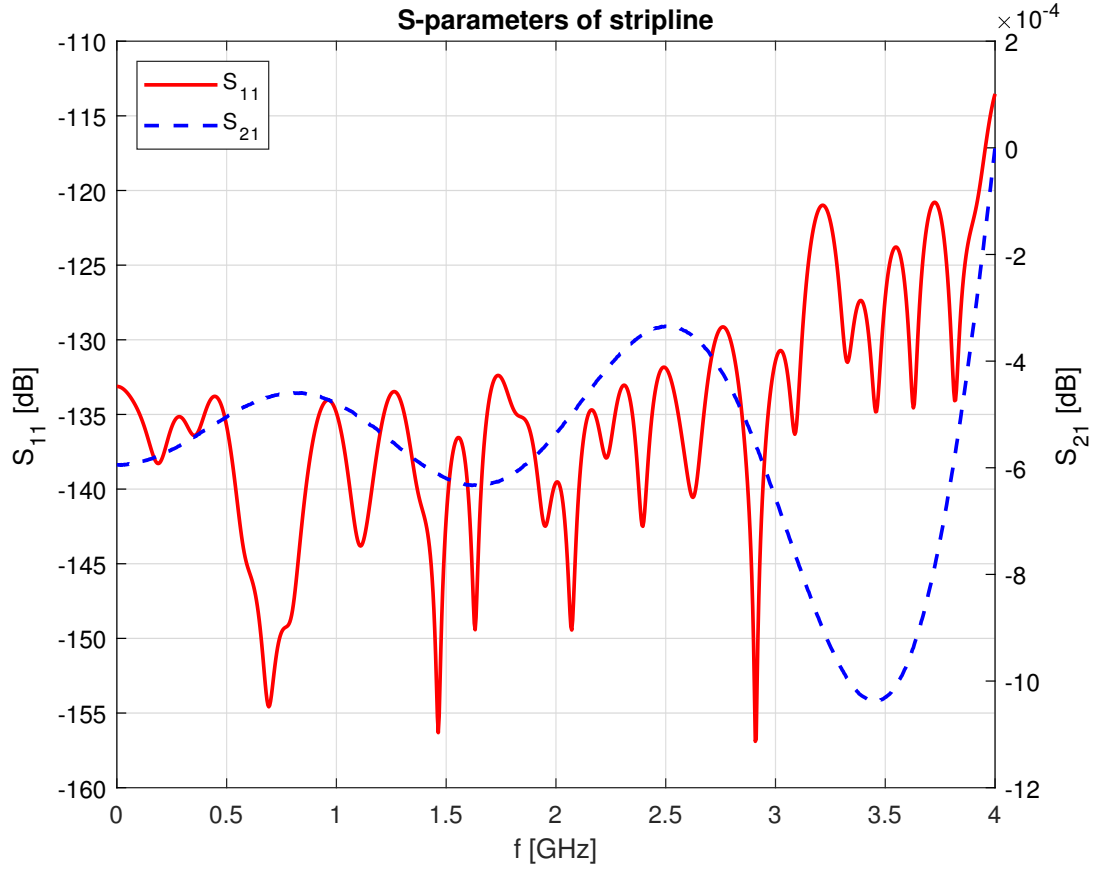


Fig. 4.3: Simulated stripline S-parameters.

4.2 Transition between coaxial line and stripline

The transition between the coaxial line and the stripline can be designed in several ways. Simply put, there are three possibilities for how to connect coaxial center conductor to stripline center conductor. The first option is to attach the circular end of the coaxial center conductor to the wider side of the stripline. The second option is to attach it to the narrow side of the stripline. The last option is to attach it to the back of the stripline. An illustration of these connections can be seen in Fig. 4.4. For simplification let's call the first variant *horizontal*, the second one *vertical*, and the third one *straight*.

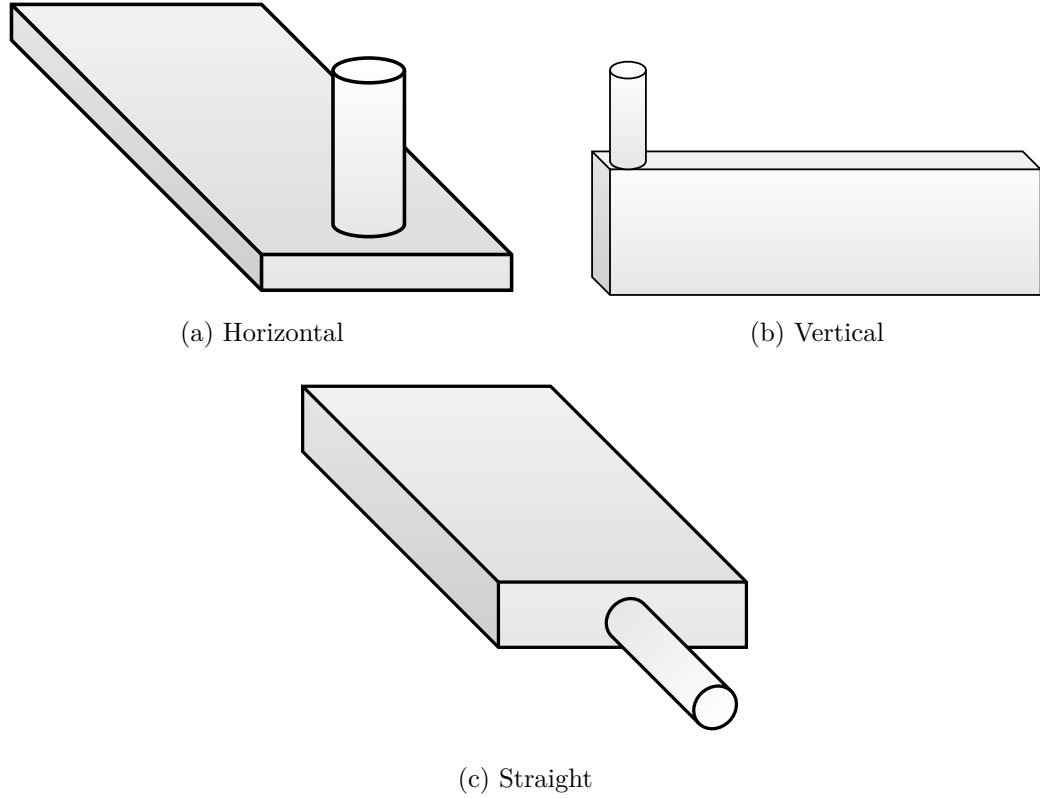


Fig. 4.4: Options of transition between coaxial line and stripline.

4.2.1 Transition design options

For the initial design, the key parameter apart from the electrical properties was also the manufacturing difficulty. The prototype had to be ready for the earliest beam time (in a few weeks) and there were not many resources and manpower to manufacture a complex design.

Straight

We considered *straight* transition between coaxial line and stripline but it is problematic. In front of the stripline from both ends, there are other components necessary for the experiment. To make this transition the stripline needs to be curved at its ends.

The first option can be seen in Fig. 4.5a. The parameter that we tried to use for tuning was the radius of the bend. Also, we changed the direct transition from stripline to coaxial line for a smoother one as can be seen in Fig. 4.5b. The electrical properties with the direct transition were decent and with the smooth transition, they were favorable. However, this design was rejected due to its complex manufacturing. There was no time to create the smooth transition and the bend was problematic as well.

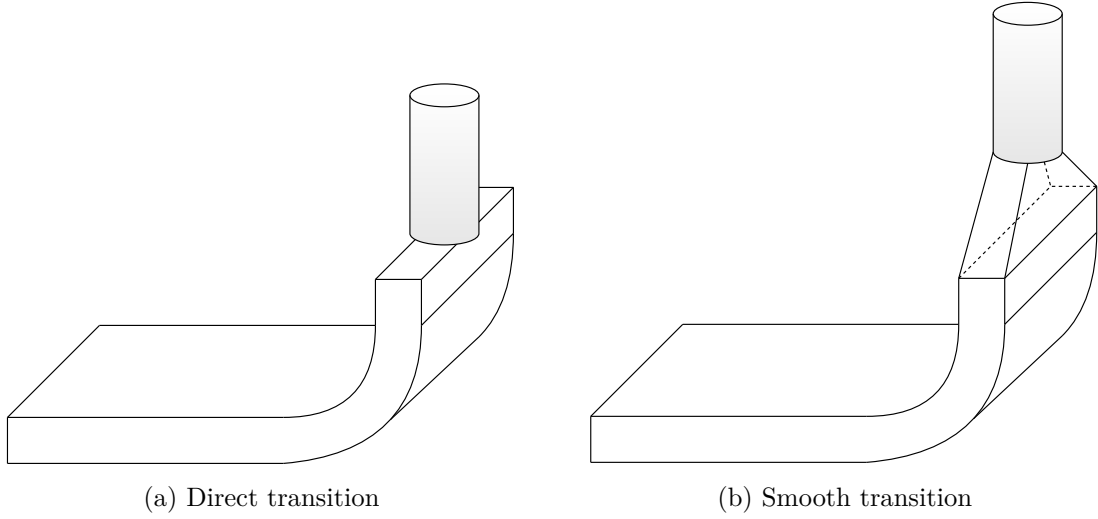


Fig. 4.5: Straight transition with horizontal bend.

We also tried a similar design but with a vertical bend. The design can be seen in Fig. 4.6. We did not consider the smooth transition variant in this case because of its manufacturing problems. Manufacturing of this shape is easier than the previous one. This strip could be cut from a metal sheet so no bending would be necessary. In the design, there are two edges as the strip is changing its direction. We also simulated this design with rounding of these edges with different diameters and chamfer with different sizes too. Nevertheless in the end after tuning all the considered designs, this one had not the smallest reflections in the considered frequency range.

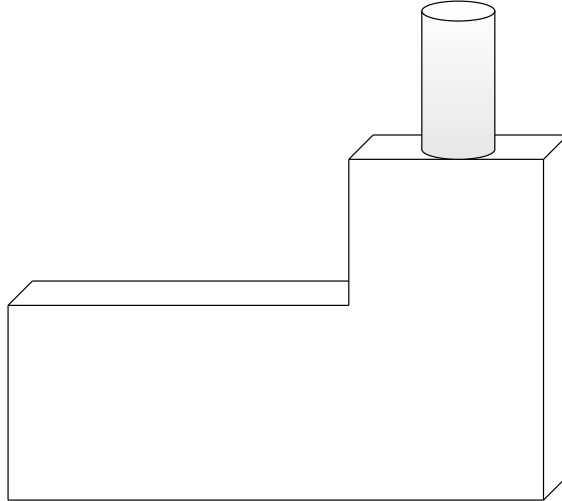


Fig. 4.6: Straight transition with vertical bend.

Vertical

We probably tried the most of parameters for tuning in the *vertical* type of design. The illustration of the design can be seen in Fig. 4.7. The first parameter we used for tuning was the position of the center coaxial conductor. We tested the horizontal and vertical position with depth of inserting of the conductor. Because of the beam, there cannot be anything sticking out of the strip on one side. For the coaxial conductor we tested inserting in the strip and different cutting options. The length of the cut had basically no effect on its electrical parameters. The same goes for the type of cut (straight, chamfer, and rounded). The best result was when the coaxial conductor was just touching the strip and was not inserted in with its edge aligned with the edge of the strip.

The second parameter was the shape and size of the round cut in the strip. The cut has a circular shape with rounding of its edge. We tested different sizes of the circular cut with different roundings of the edge. These parameters could be used for pretty effective tuning.

The third parameter was the shape of the bottom edge under the coaxial conductor. We tried different sizes of rounding or chamfer. We also tried creating a heel sticking out. However, these changes made the reflections only bigger.

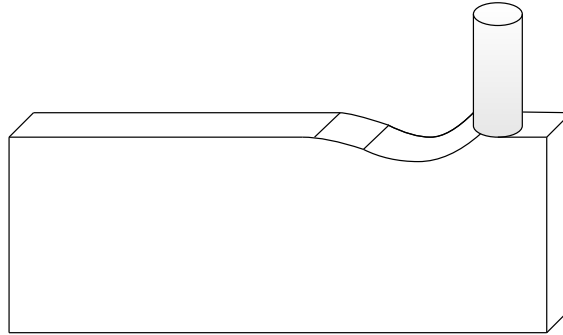


Fig. 4.7: Vertical transition.

Horizontal

The last option is probably the easiest one for manufacturing. It is a direct connection of the coaxial center conductor to the wider side of the strip. During the tuning, we tested the position of the coaxial conductor and the size and shape of the cuts. The illustration of the design can be seen in Fig. 4.8. The position of the coaxial conductor seems to be the best when the conductor is only touching the strip and is not inserted in. And when the coaxial conductor has half of its diameter on the strip and half of its diameter outside as can be seen in Fig. 4.9.

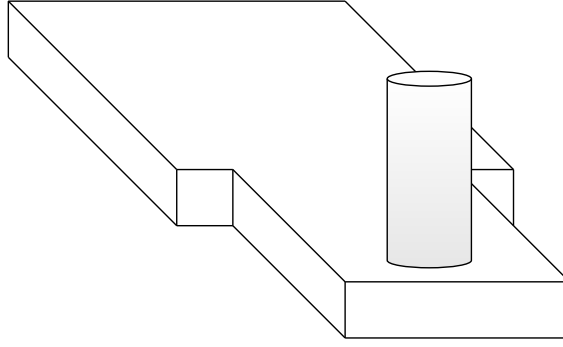


Fig. 4.8: Horizontal transition.

The size of the cuts on the side of the strip has a great influence on the electrical properties and can be effectively used for tuning. In the end, we have chosen *horizontal* design for the initial design because of its simple manufacturing and its good simulation results compared to the rest of the designs.

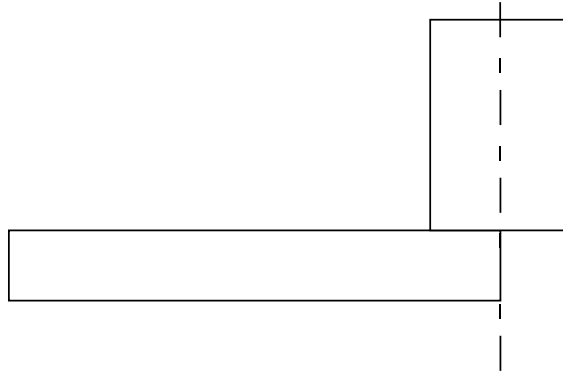


Fig. 4.9: Position of the coaxial center conductor to the strip.

Comparison

A comparison of roughly tuned several transition options can be seen in Fig. 4.10. The design *A* is the straight design with a horizontal bend as can be seen in Fig. 4.5a. It has the lowest reflections of all designs, but bending the stripline is difficult for manufacturing thus this design was refused for the initial design. The design *B* is the straight design with vertical bend as can be seen in Fig. 4.6 but with rounded edges. Its results are similar to design *E* which is also straight design with a vertical bend but with beveled edges instead of rounded ones. These two designs have bigger reflections on the lower frequencies but lower reflections on higher frequencies compared to design *C*. Design *C* is horizontal as can be seen in

Fig. 4.8. Design *D* is the vertical design as can be seen in Fig. 4.7 and it has the biggest reflections.

The lowest reflections are achieved with bent designs but their manufacturing is too complicated for the initial design. Thus the horizontal design was selected for the initial design.

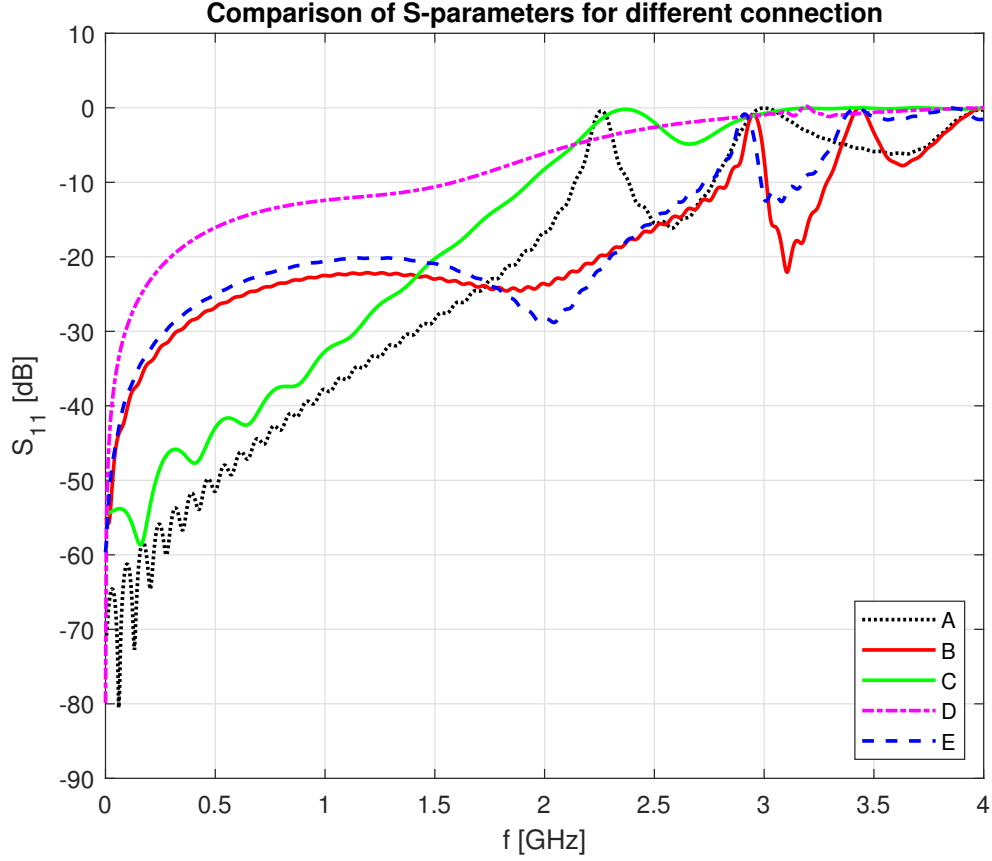


Fig. 4.10: Comparison of S-parameters for different connections.

4.2.2 Coaxial line

Because the stripline needs to be in a vacuum we created a coaxial line with a specific shape. The coaxial line is made from three pieces. The first piece is a coaxial line with a vacuum dielectric and is connected to the strip. The other end of this line ends in a hole in the top ground conductor and is connected there to the second piece of the coaxial line.

The second piece is a coaxial line with teflon dielectric. The center conductor of this piece is slightly smaller and the outer conductor diameter is slightly bigger than the hole in the top ground conductor. Thus this teflon can seal the hole.

The third piece is also a coaxial line with a teflon dielectric. The center conductor of this piece is bigger than the previous one. Thus this prevents the center conductor from falling into the rectangular tube that serves as the outer ground conductor of the stripline.

We calculated every piece in CST Studio coaxial line calculator and then fine-tuned them in simulations. The coaxial line can also be calculated by hand with the use of equation 2.1.

This coaxial line provides not only signal feed for the stripline but also works as a seal and holds the stripline in a desired position. We calculated the diameters according to ISO-KF 40 tube that serves as the outer conductor of the coaxial line and was requested due to mechanical connection options. An illustration of the design can be seen in Fig. 4.11.

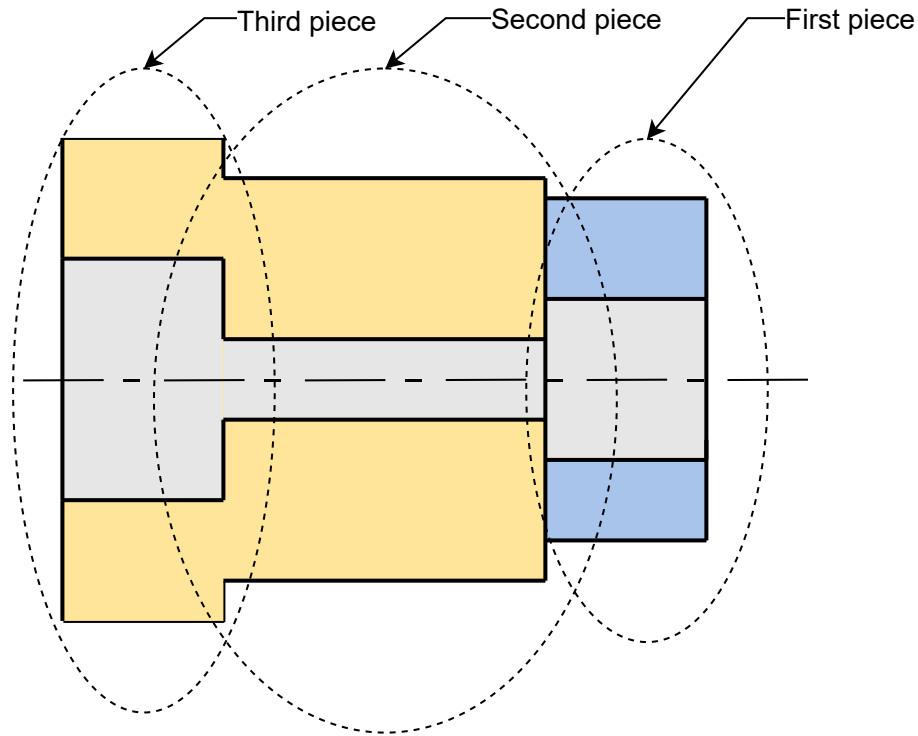


Fig. 4.11: Coaxial line design.

Simulation results of the coaxial line can be seen in Fig. 4.12. The S_{21} parameter is good even in higher frequencies. But the S_{11} parameter is bad, especially on higher frequencies. In the band considered for the initial design, the parameters are not ideal but sufficient and its mechanical properties are really needed for the design. But some changes are needed for the final design.

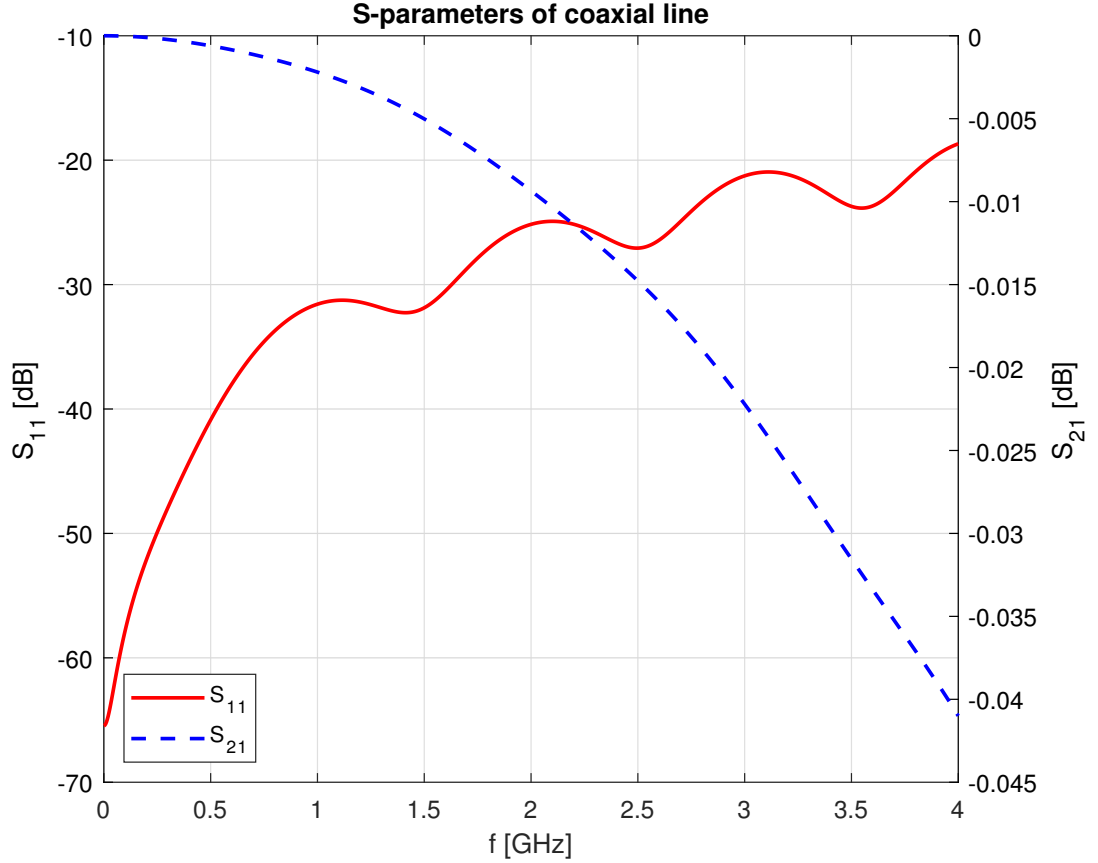


Fig. 4.12: Coaxial line S-parameters.

4.2.3 Simulation of the transition

At the beginning of the tuning process, we modeled only half of the transmission line with a shorter length for faster simulation. After tuning all considered designs, we chose the *horizontal* transition because of its best result and easy manufacturing.

The design contained the stripline, the first section of the coaxial line mentioned before at subsection 4.2.2, and the conducting wall at the end of the stripline. The design modeled in CST Studio can be seen in Fig. 4.13. There is a hole in the wall because the beam needs to be able to get in and out of the structure.

The best achieved results can be seen in Fig. 4.14. S_{11} parameter is below -50 dB in considered frequency range and S_{21} is above -0.5 dB. Port 3 is the hole in the wall thus parameters S_{31} and S_{32} are related to it. They are also below -50 dB in the considered frequency range. Over all the parameters are favorable and this design is also the easiest to manufacture.

The next step was to change the shape a bit so it could be assembled and disassembled if necessary. One requirement was to have an option of disassembling

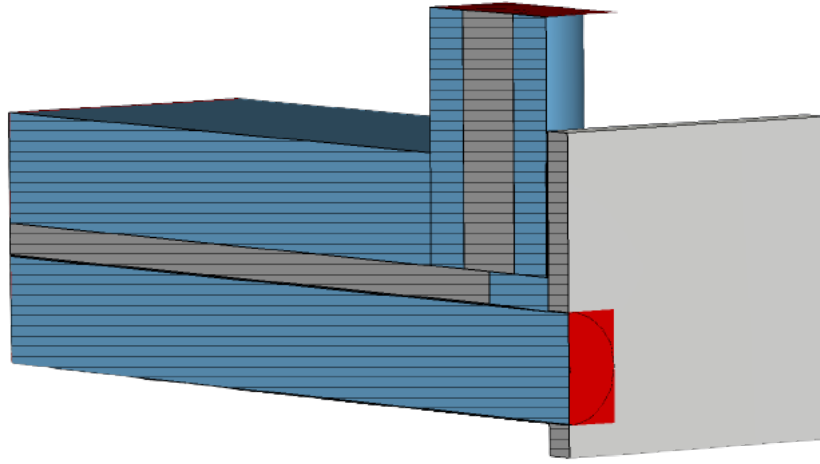


Fig. 4.13: Horizontal design in CST Studio.

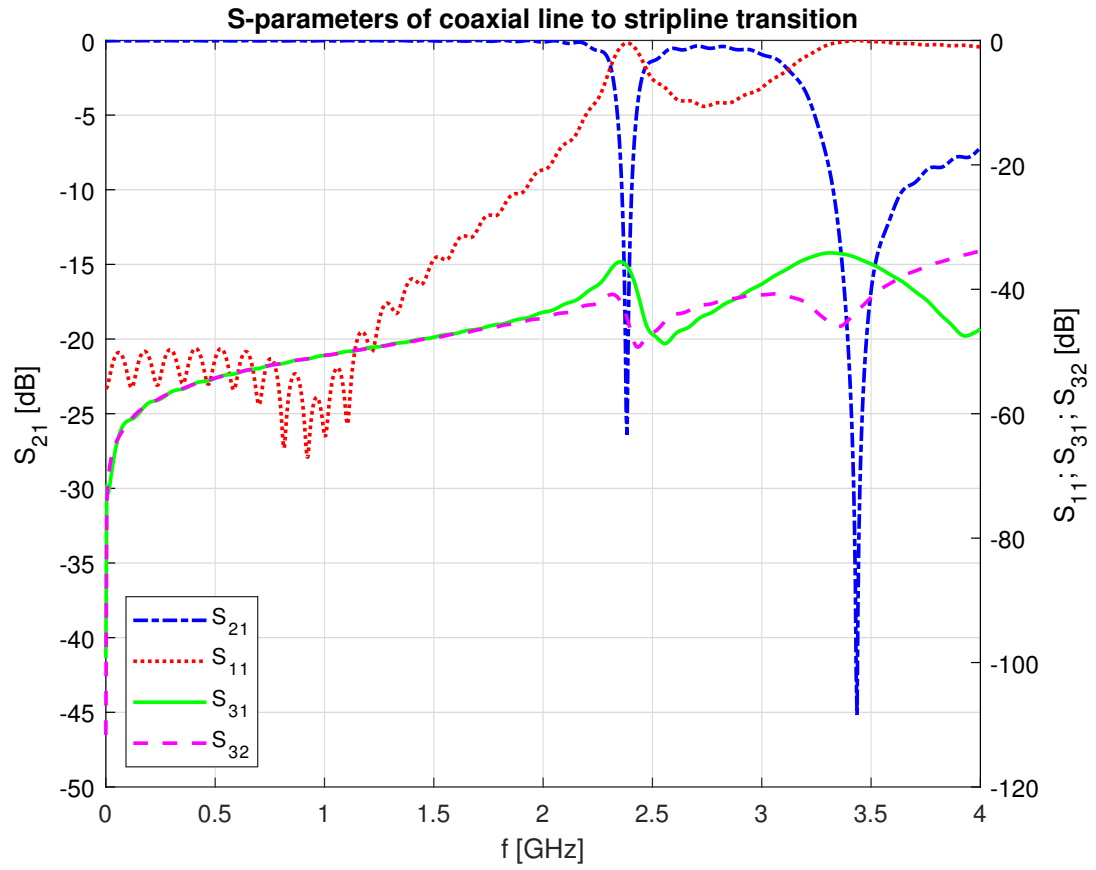


Fig. 4.14: The best achieved results of the transition.

the connection of the coaxial line and the stripline. The easiest solution is to change the position of the coaxial center conductor and make a thread at the end. Then make a hole with a thread to the strip and screw the coaxial center conductor to the strip.

We have created two such designs. The first one by adding half a circle to the end of the strip with the same diameter as the coaxial center conductor. In the strip, there is a hole with a thread and at the end of the coaxial center conductor, there is an extension with a thread. An illustration of this connection can be seen in Fig. 4.15a.

The second design is created by simply moving the coaxial line closer to the center of the strip. Then the connection of the coaxial center conductor to the strip is the same as in the previous design. An illustration of this connection can be seen in Fig. 4.15b. Both of these designs were simulated and the cuts on the sides were tuned again. As can be seen in Fig. 4.15 in the first case the results were best without the cuts.

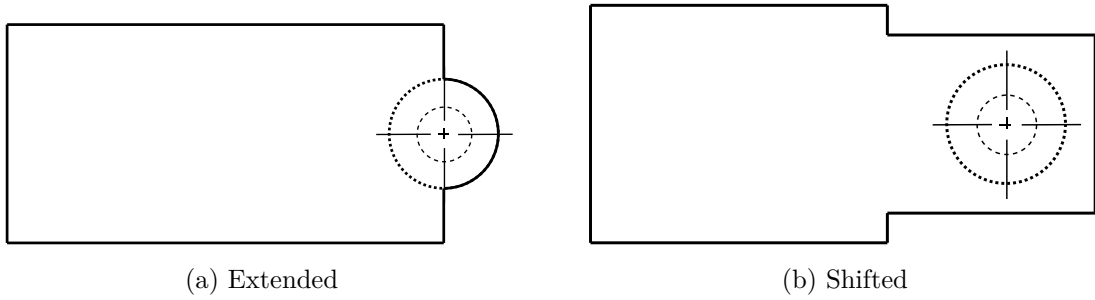


Fig. 4.15: New transition design that can be disassembled.

The S-parameters can be seen in Fig. 4.16. Curves labeled with a are for the extended design and curves labeled with b are for the shifted one. S_{21} is good in both cases but slightly better with the extended design. S_{31} is basically the same. S_{32} is also really similar but slightly better with the shifted design. The main difference is in S_{11} parameter. At lower frequencies, the shifted design is slightly better, and at higher frequencies, the extended one is better. Both designs were simulated in full form with real materials, full length, and complete coaxial line.

We added supports to the sides of the stripline. The stripline will be 2 m long and the supports supposed to prevent it from bending. The illustration of the supports can be seen in Fig. 4.17. The supports are from both sides spaced every 200 mm and they are made of teflon. Their diameter is the same as the thickness of the strip and their length is equal to the difference of half of the substrate width and half of the strip width. To further prevent bending the strip is directed with its narrower side to the ground.

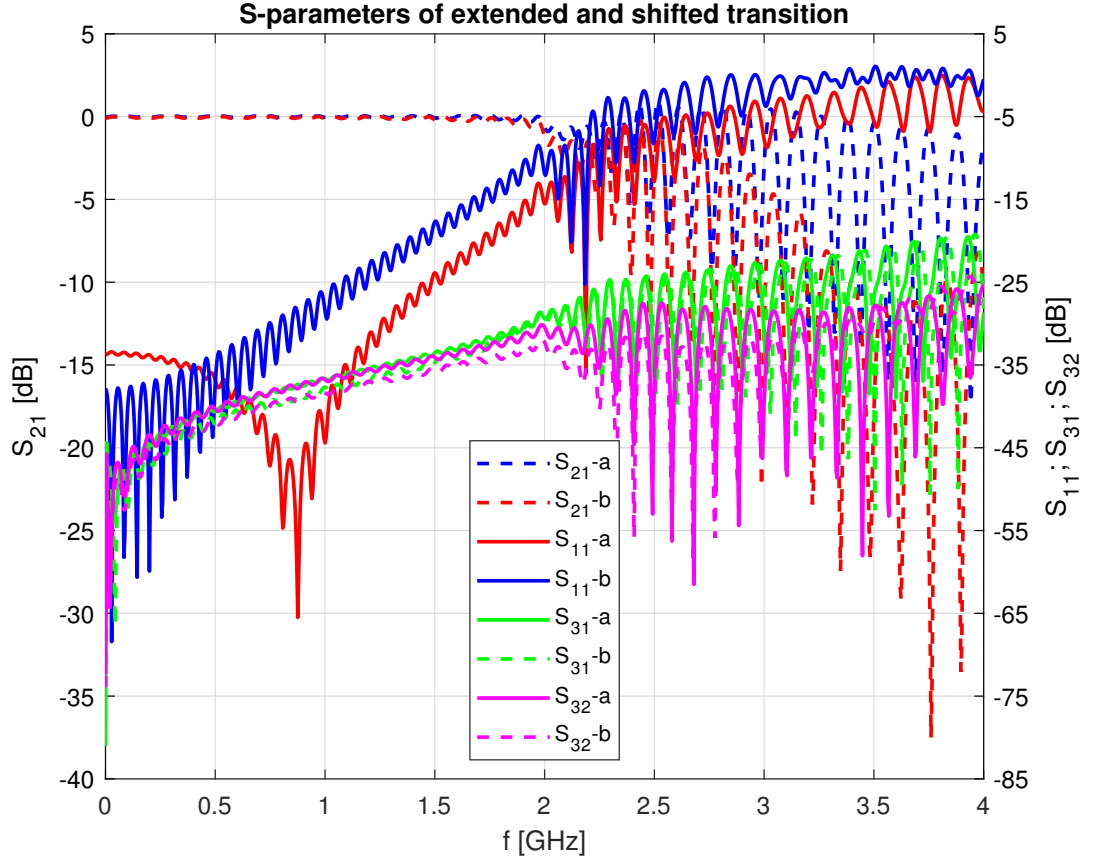


Fig. 4.16: Comparison of S-parameters of extended (a) and shifted (b) design.

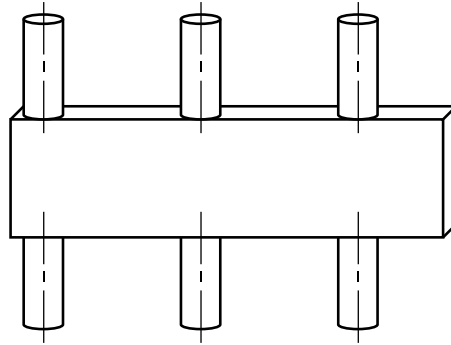


Fig. 4.17: Stripline with supports.

We simulated the effect of the supports on the S-parameters with the extended design at full length and only with the first part of the coaxial line. The S-parameters can be seen in Fig. 4.18. The curves labeled with n are without the supports and curves labeled with y are with the supports. The supports have basically no effect

on the S-parameters. S_{21} , S_{31} and S_{32} is basically without change. The only small difference is with S_{11} parameter. With the supports, the reflections are slightly higher but this difference is minor.

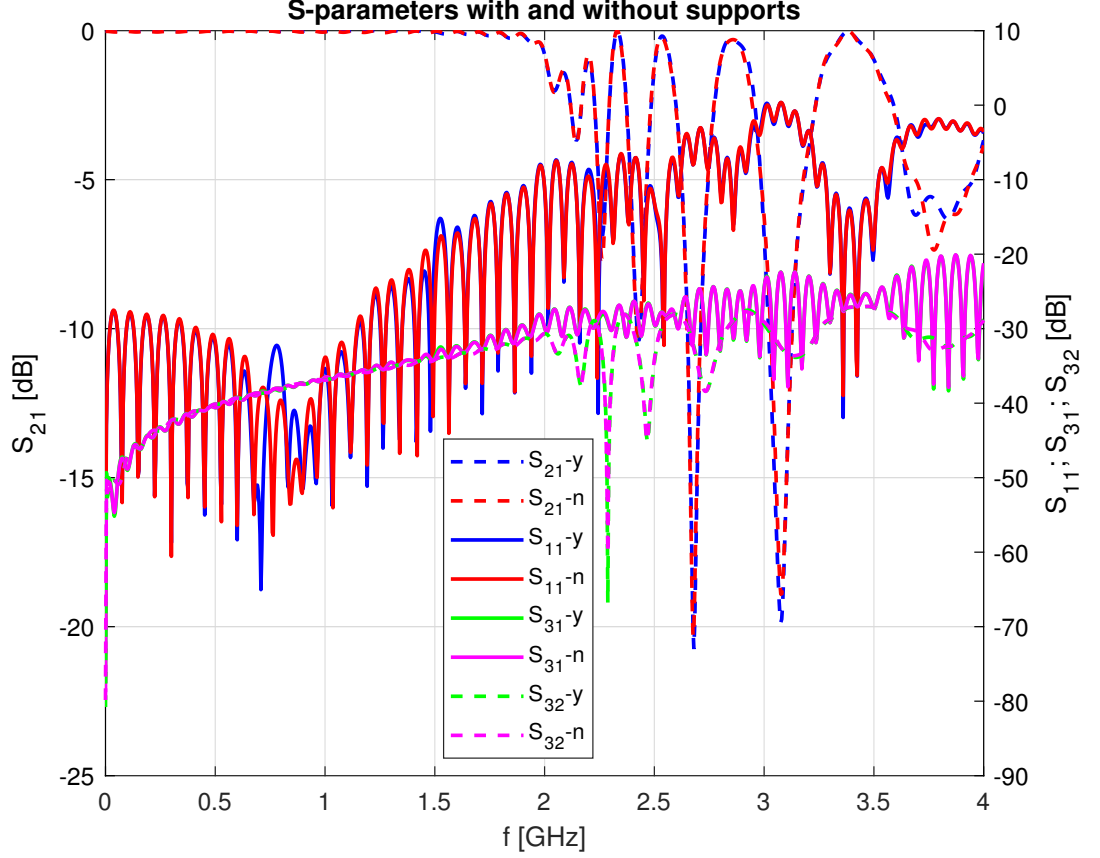


Fig. 4.18: Comparison of S-parameters with (y) and without (n) supports.

Finally, we simulated both designs at full length. Supports and the full coaxial line were added and PEC was changed to aluminum. Results can be seen in Fig. 4.19. S_{32} and S_{31} parameters are slightly better with the extended design. S_{21} parameter is basically the same for both designs. S_{11} parameters are slightly better with the shifted design.

In the end, for the final construction, we have chosen the shifted design. The main reason is better S_{11} parameters and easier manufacturing. The next step is to model everything in 3D CAD (Computer Aided Design) software, create technical drawings, and construct the whole design.

S-parameters of extended and shifted full length design

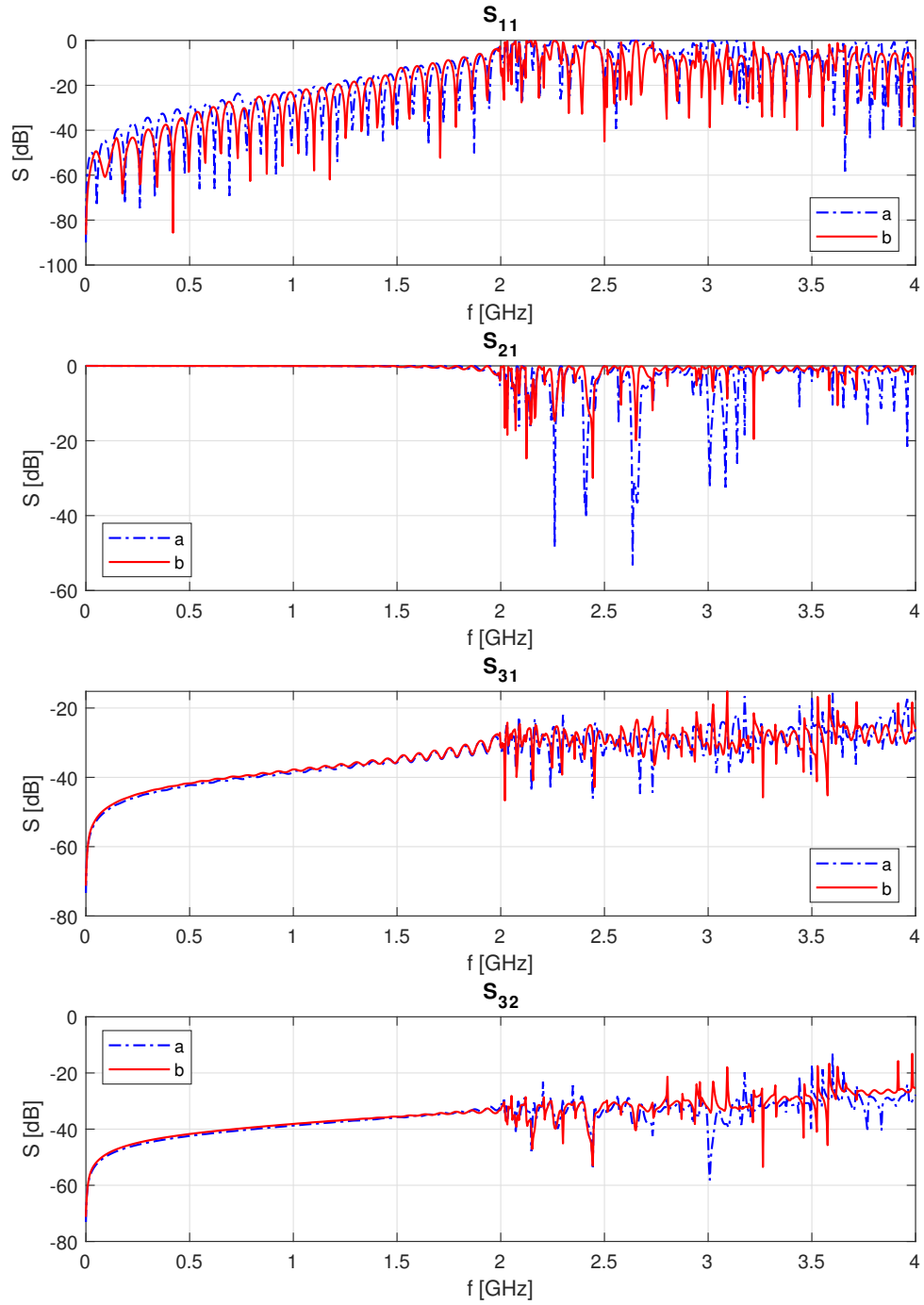


Fig. 4.19: Comparison of S-parameters of extended (a) and shifted (b) design.

5 Construction of the Initial Design

For the construction of the stripline, the most challenging part was to design it with the option of disassembling it while keeping it vacuum-tight. We modeled every part in CAD software Autodesk Inventor. We also made technical drawings for manufacturing the design using the same CAD software.

The first step in the assembly of the stripline is to attach the supports and the first part of the coaxial line to the strip itself as can be seen in figure 5.1. The supports will be glued to the strip and the coaxial line is attached by a screw. Supports were added only to one side of the strip. Their effect on the S-parameters was minor and it would be easy to add them later if necessary.

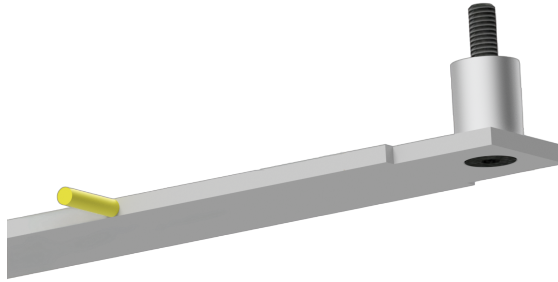


Fig. 5.1: Attaching supports and the first part of the coaxial line to the strip.

The next step is to assemble the coaxial line described in subsection 4.2.2. The vacuum part of the coaxial line is already attached to the strip. Now the inner coaxial conductor is inserted into the teflon dielectric. Then the teflon is inserted into the first conducting tube that serves as the outer conductor. After this, the whole thing is inserted into ISO KF-40 tube. On the top of it, there is a centering ring, ISO KF-40 flange with connector, and it is tightened together by clamp as can be seen in Fig. 5.2.

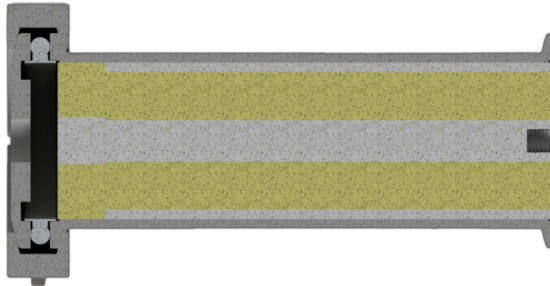


Fig. 5.2: Coaxial line construction.

Then ISO K-100 flange is welded to the aluminum tube that serves as a ground plane. In the flange, there is a rectangular cut for attaching it to the tube and for the beam so it can get inside of the structure as can be seen in Fig. 5.3. Also on the top of the aluminum tube, there are drilled holes for the coaxial line.

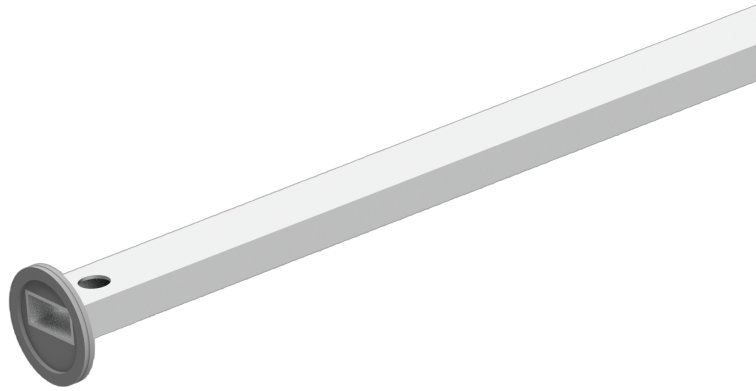


Fig. 5.3: Ground plane construction.

The last step is to attach all previously mentioned pieces together. The strip is inserted into the ground plane and the coaxial line is attached by a screw in the strip. Between the coaxial line and the tube, there is an O-ring for keeping a vacuum-tight design. The coaxial line is attached to the tube by a holder and screws as can be seen in Fig. 5.4

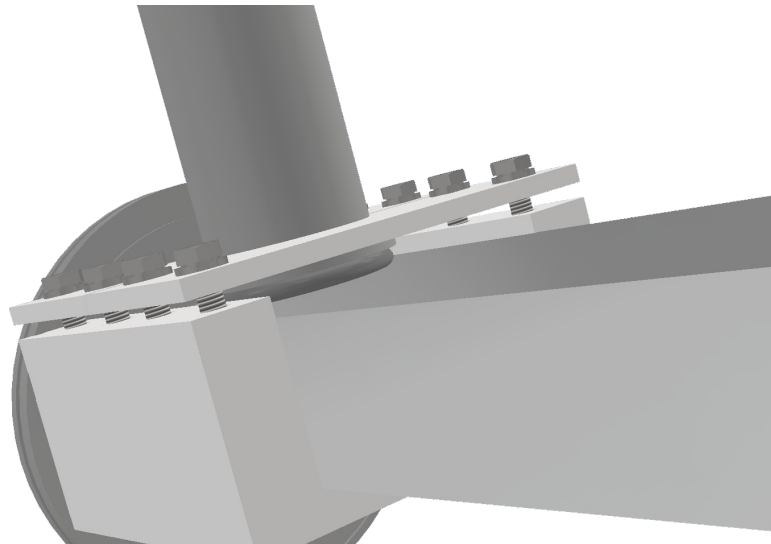


Fig. 5.4: Stripline assembly.

At the end of the tube, there is a wall attached with 2 headless screws and 4 nuts. In the wall, there is a hole for the beam. The wall attachment can be seen in Fig. 5.5.

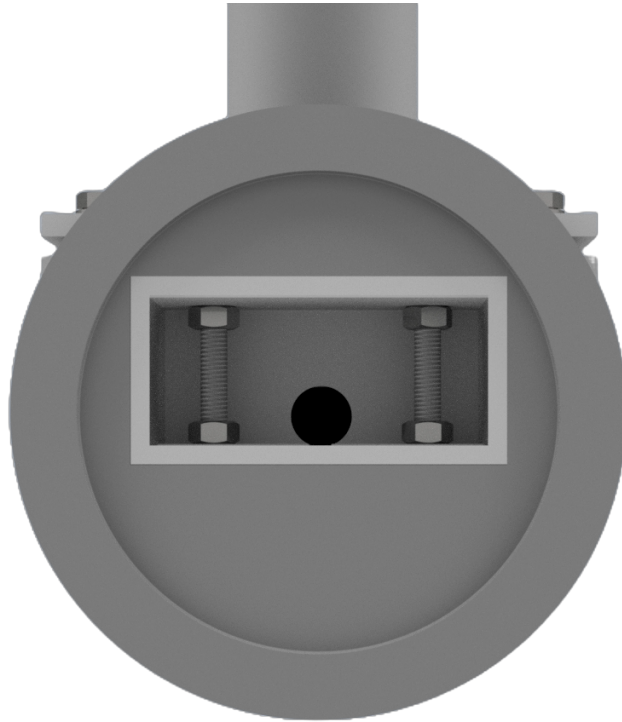


Fig. 5.5: Wall attachment.

6 Final design

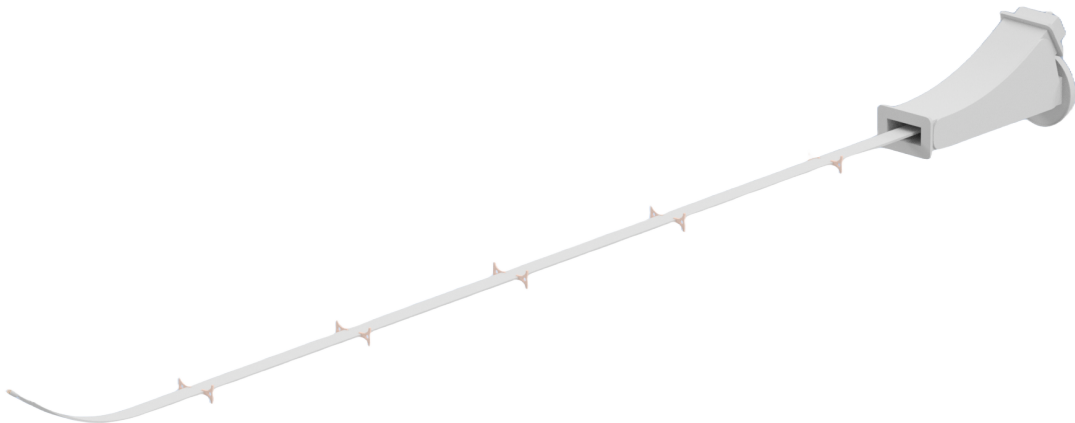
For the final design, some adjustments were needed. The most important was to change the transition from coaxial line to stripline to increase the frequency range. Also, we changed the design of the supports.

6.1 First version

For the first version of the final design, we used the *straight* connection with *horizontal* bend and smooth transition as can be seen in Fig. 4.5b. For this design, we adjusted the supports's design and tuned the transition to obtain the lowest reflections. We modeled the full design partially in Autodesk Inventor and partially in CST studio. The model fo the full design can be seen in the following Fig. 6.1.



(a) Outside



(b) Inside

Fig. 6.1: First version of the final design.

6.1.1 First version - supports

We changed the support design to provide better positioning of the stripline inner conductor. We tested several geometries. The illustration of the most suitable shapes can be seen in Fig. 6.2.

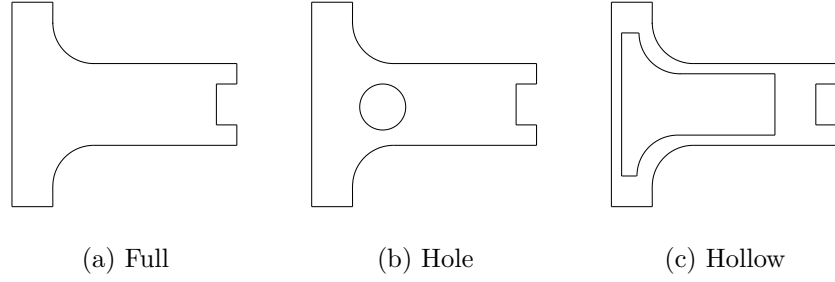


Fig. 6.2: New support design.

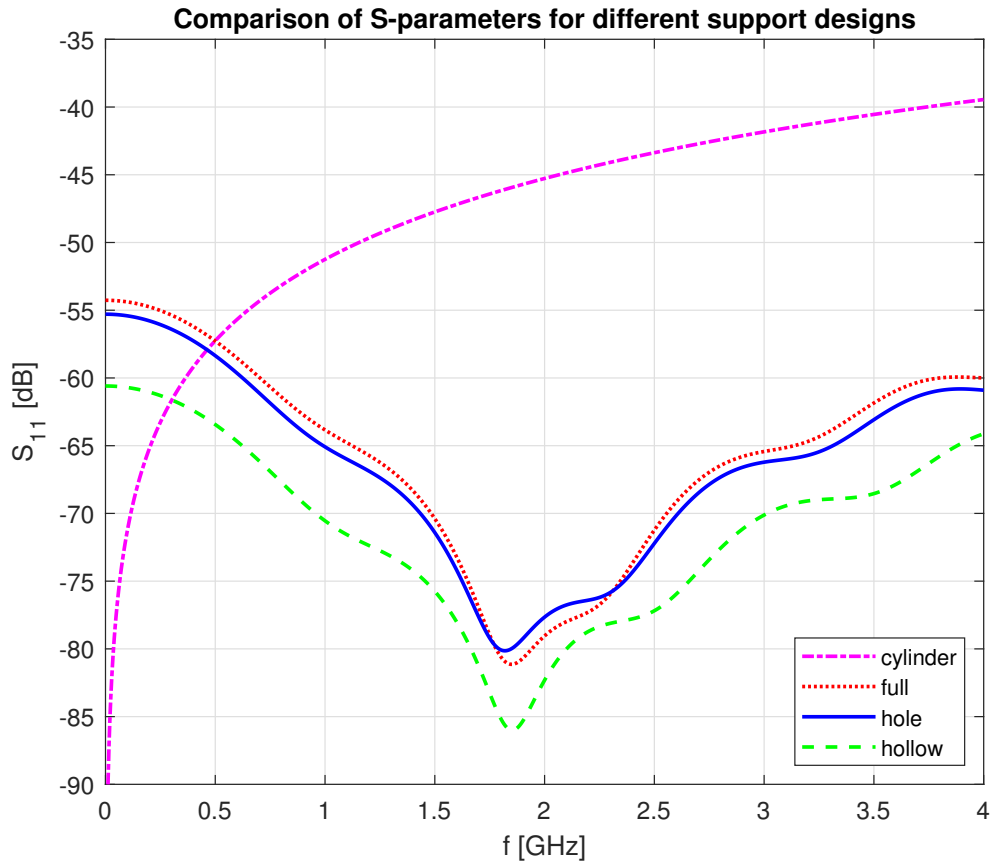


Fig. 6.3: Support design comparison.

In Fig. 6.3, S-parameters comparison can be seen. We simulated 4 mm long stripline with the specific support. The thickness of the supports was the same as the length of the stripline. The previous cylindrical supports have the biggest reflections and provide the least support. The lowest S_{11} parameter can be achieved with the *Hollow* design. But the strength of this design was not sufficient and the support would be deformed. This is why we chose the *Hole* design that has sufficient mechanical properties and reflections. The supports were placed every 400 mm from both sides of the strip as can be seen in Fig. 6.1b.

The comparison of S-parameters of design with and without support can be seen in Fig. 6.4. The difference between design with and without supports is minor. With this design, we managed to obtain a higher frequency range up to almost 2 GHz. Around this frequency, resonances start to appear in the S-parameters.

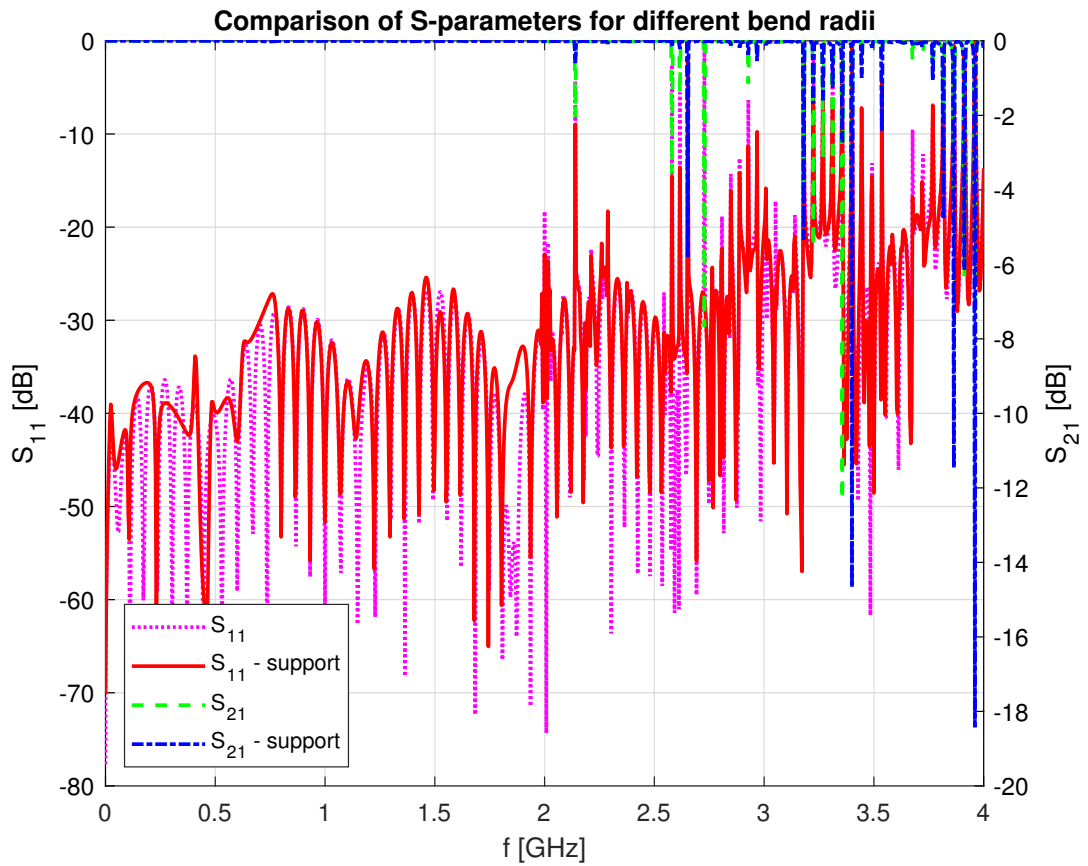


Fig. 6.4: S-parameters of the first version of the final design with and without supports.

6.1.2 First version - increasing frequency range

In Fig. 6.4 it can be seen that on frequencies higher than 2 GHz peaks in reflections appear. We tried to remove the peaks by changing the bend radius. By changing the radius, the position and peak value of the peaks can be changed, but this tuning is not very effective. In Fig. 6.5 there can be seen a comparison of S-parameters for different bending radii. We did the simulation only for the last bent piece of the whole apparatus without the transition region. At first, it seemed that we could remove the peaks by increasing the radius. but when we added more points to the simulation new more narrow peaks started to appear.

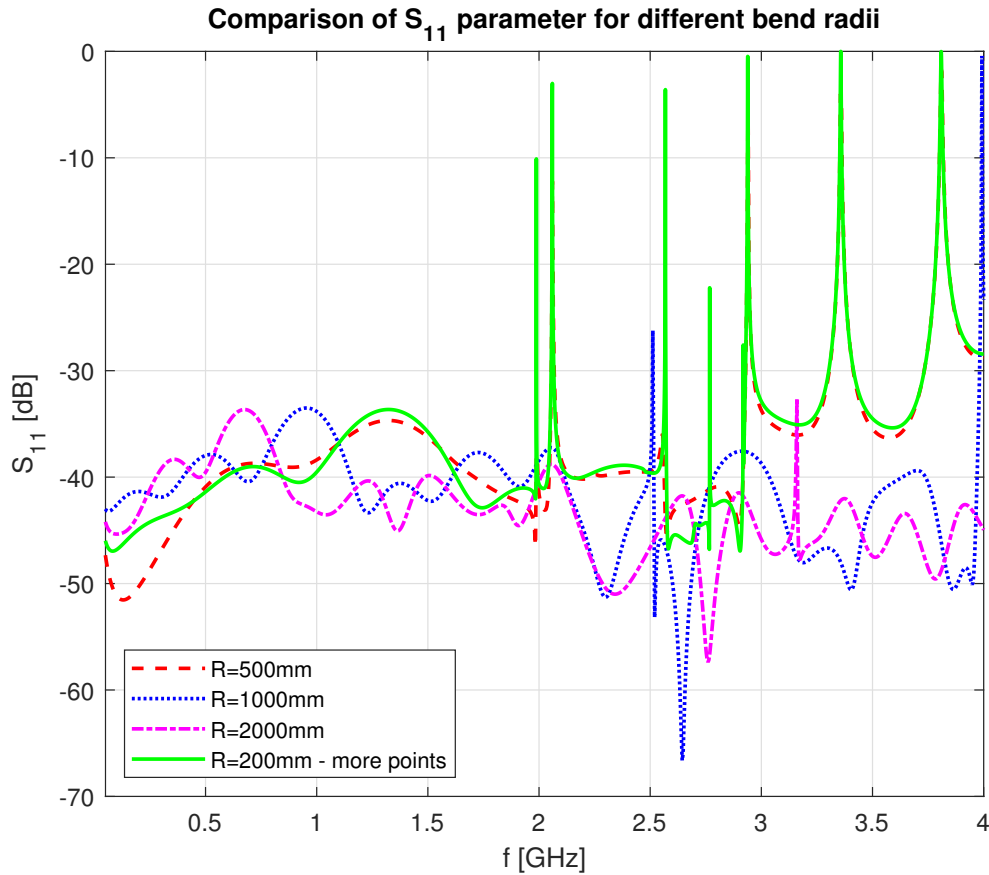


Fig. 6.5: Comparison of S_{11} parameter for different bending radius for the first version of the final design.

The start frequency of the peaks is dependent on the size of the stripline. A higher frequency range can be achieved by decreasing the dimensions of the whole design. If we decrease the dimensions by half we can obtain a frequency range up to 4 GHz. A comparison of normal size, double size, and half size can be seen in Fig. 6.6. The comparison was done for the bent piece with the transition to coaxial

line with bending radius 250 mm. Unfortunately, this approach cannot be used due to not enough space for the beam. The sizes of the outer conductor cannot be decreased.

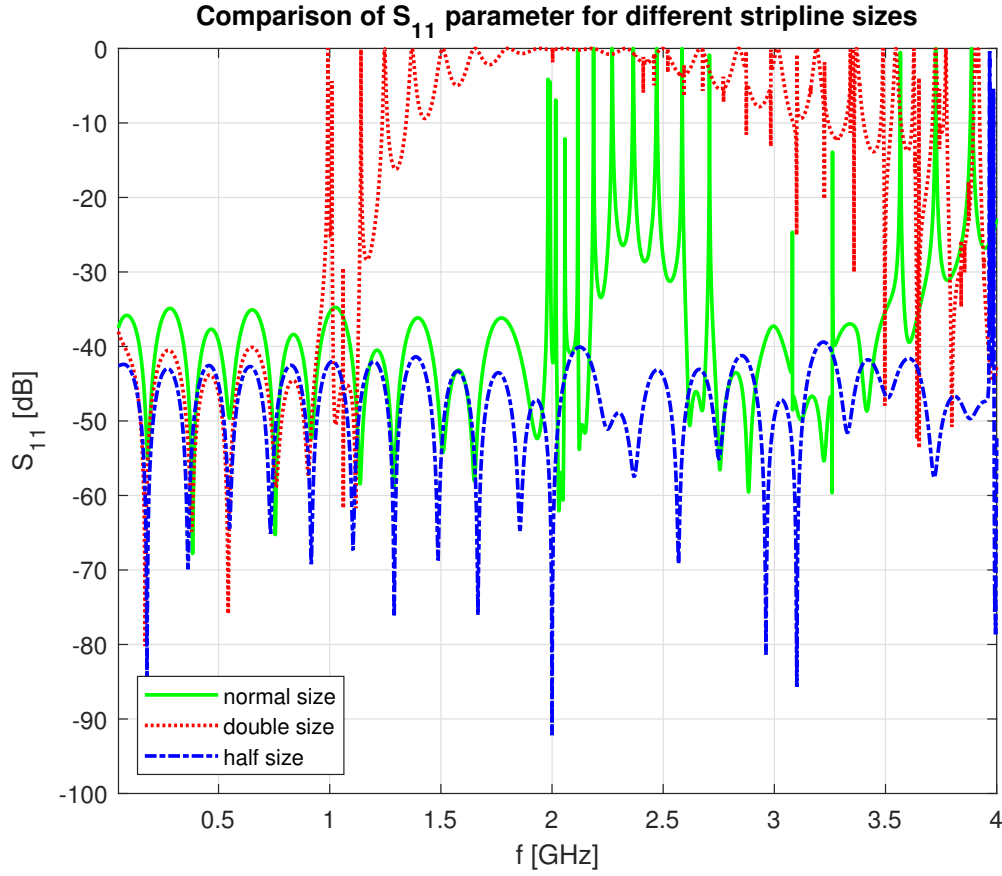


Fig. 6.6: Comparison of S_{11} parameter for different sizes of the first version of the final design.

Another thing we tested was the bending direction. Instead of bending the stripline in *horizontal* way we tried to bend it in *vertical* way. Illustration of this bend can be seen in Fig. 4.6. The differences were that the bend was not direct 90° but a smoother one with bending radius 250 mm. A comparison of the results can be seen in Fig. 6.7. With this adjustment, we managed to increase the frequency range by 0.5 GHz. Thus the achieved frequency range for this design is from 0 GHz up to 2.5 GHz.

This frequency range is sufficient for most of the measurements. But there might be a possibility for the need for a higher frequency range. Due to this reason, we decided to try to develop a new design with even higher frequency range.

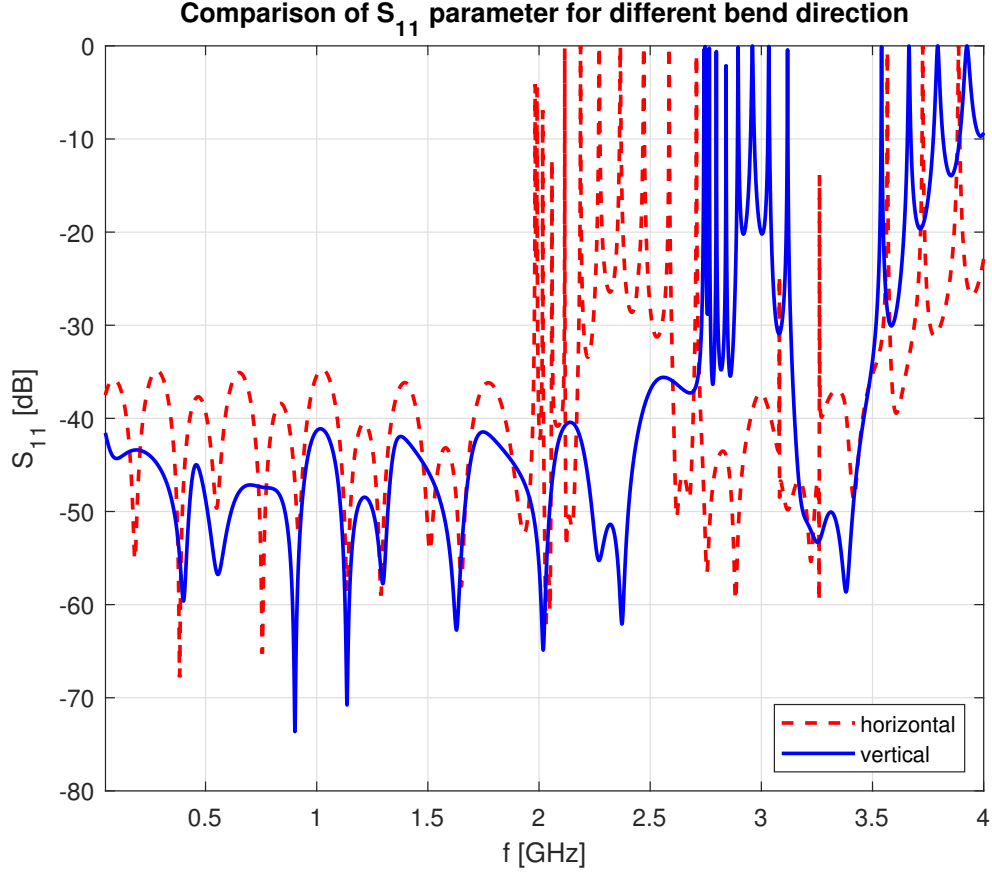


Fig. 6.7: Comparison of S_{11} parameter for different bending directions for the first version of the final design.

6.2 Second version

We designed the second version to obtain even higher frequency range. In this case, the center conductor of the stripline goes straight to the center conductor of the coaxial line. The stripline and the transition are straight. The coaxial line is then bent to reach the feedthrough that needs to be on the side of the design. Otherwise, it would block the beam.

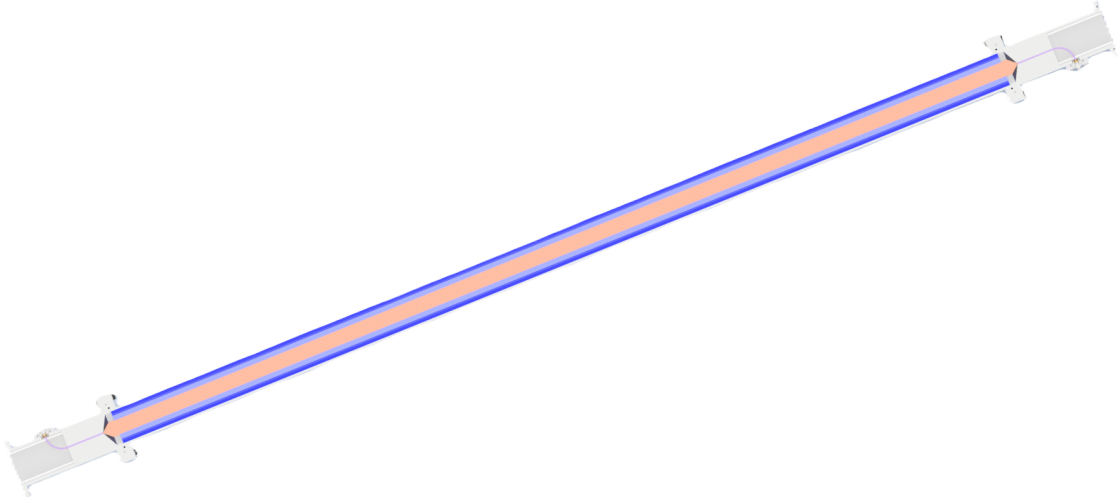
The problematic part of this design is the limited space for the coaxial line. Between the center of the design to the edge of the hole for the beam, there is 2 mm distance. That limits us to coaxial lines with diameters smaller than 4 mm. This brings problems with mechanical connections and manufacturing and the coaxial cable needs to be able to transmit enough power (about 100 W).

For the coaxial cable, we chose semi-rigid cable RG402. The cable is without outer insulation. That allows bigger-sized cable to be used and also it provides

a good and easy connection between the ground of the coaxial line and the outer ground conductor of the apparatus and it is suitable for our frequency range and transmitted power. As a connector, we decided to use SMA (SubMiniature version A) instead of BNC (Bayonet Neill Concelman) which was considered for the first version because of its higher frequency range and smaller sizes. The selected coaxial cable is simply too small for BNC connector. We designed the full model in Autodesk Inventor and can be seen in the following Fig. 6.8.



(a) Outside



(b) Inside

Fig. 6.8: Second version of the final design.

6.2.1 Second version - supports

For this design, we improved supports. Instead of using multiple short supports, we decided to use one big support along the whole stripline center conductor. It provides better mechanical support and the reflections are also smaller. We also tried to put the supports zigzag along the stripline but the results with one solid piece were better. The comparison of S_{11} parameter for *Hole* and *Hollow* support shape (that can be seen in Fig. 6.2b and 6.2c) for one solid piece or multiple of them can be seen on Fig. 6.9.

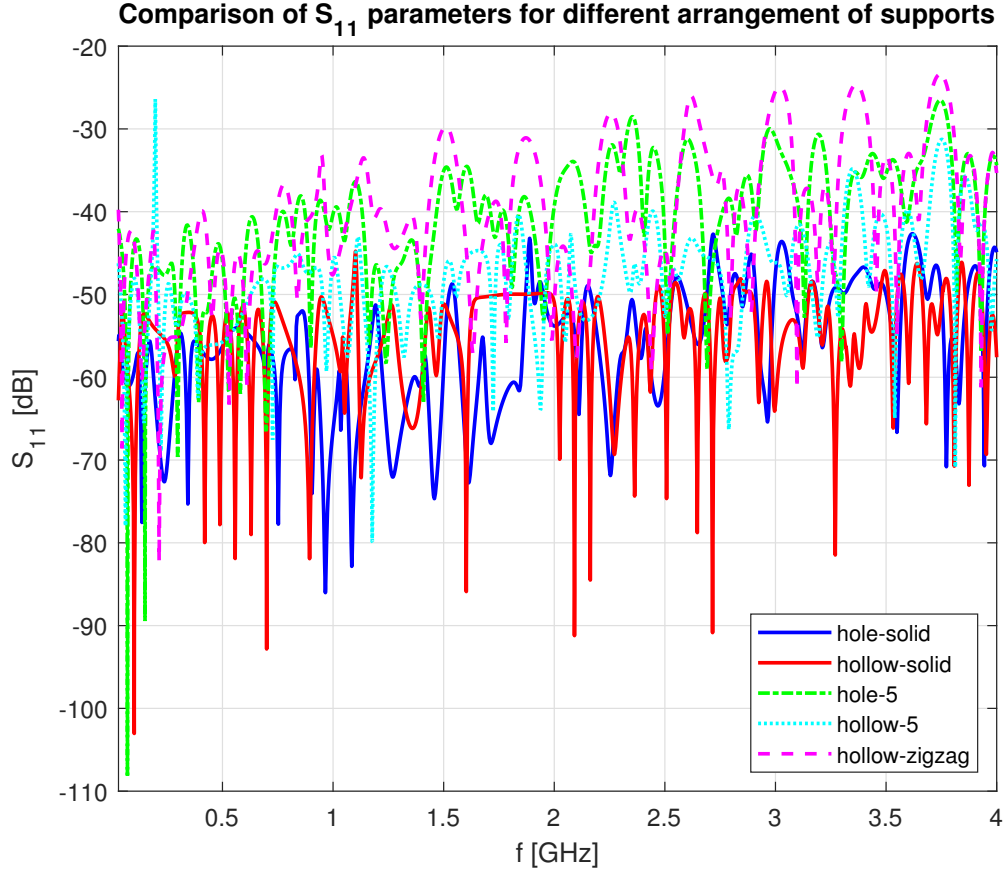


Fig. 6.9: Comparison of S_{11} parameters for different arrangement of supports and their shape.

The simulation was done for 2m long stripline with supports but without any transition to the coaxial line. As can be seen in Fig. 6.9 the zigzag arrangement has the highest reflections. We achieved the smallest reflections with only one solid block of support.

But for the use of one solid piece of support, we needed to change the shape of the support. Manufacturing 2m long support with a hole inside is not possible in

our conditions. Thus we tested new shapes that can be seen in the following Fig. 6.10.

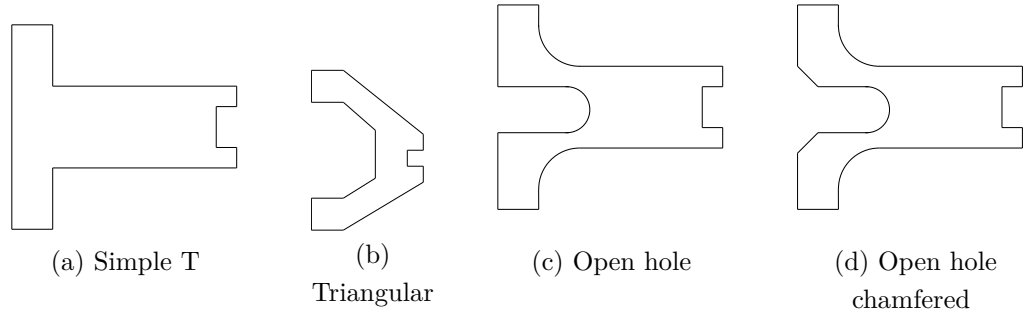


Fig. 6.10: Solid support options.

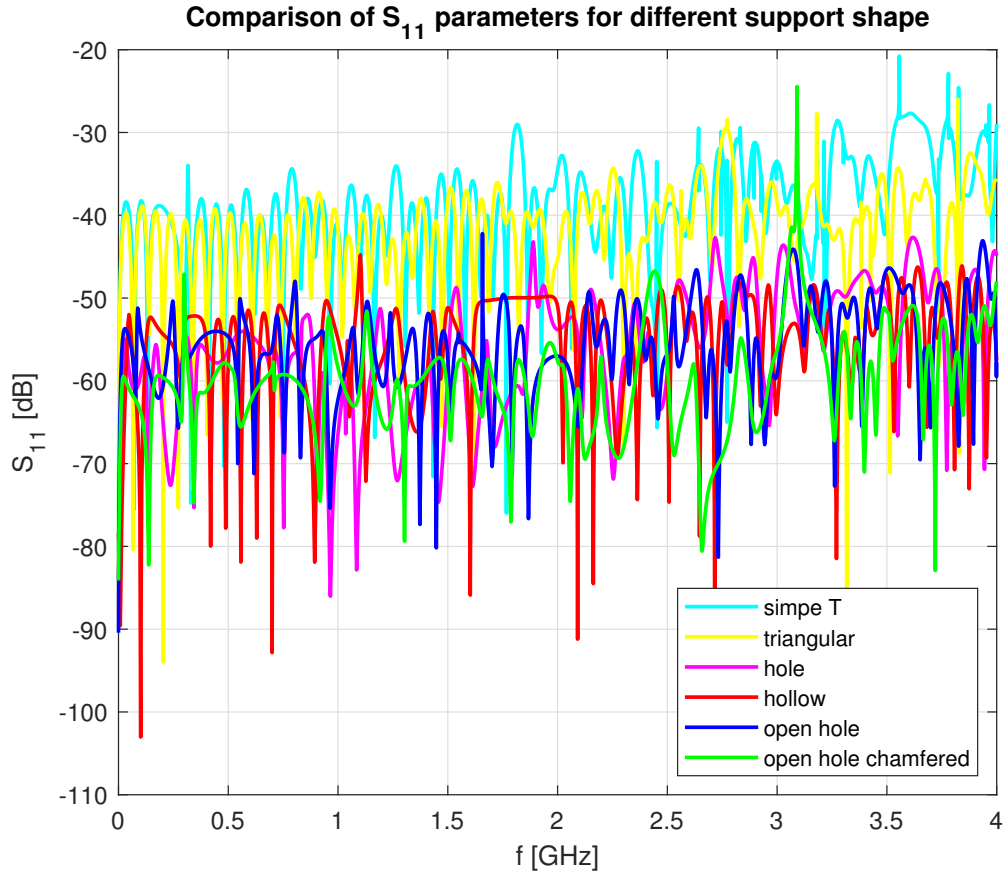


Fig. 6.11: Comparison of S_{11} parameter for different support shapes.

The comparison of S_{11} for the new and also some previous support shapes can

be seen in Fig. 6.11. The lowest reflections were achieved with *Open hole* shape that can be seen in Fig. 6.10c and was chosen for the final design.

6.2.2 Second version - simulation results

Initially, we designed the apparatus with only one hole in the bottom part for the beam. But the reflections were high. When we added a second hole of the same size to the top part of the design to be symmetrical and the reflections were significantly decreased. The comparison of the design with one and with two holes can be seen in Fig. 6.12.

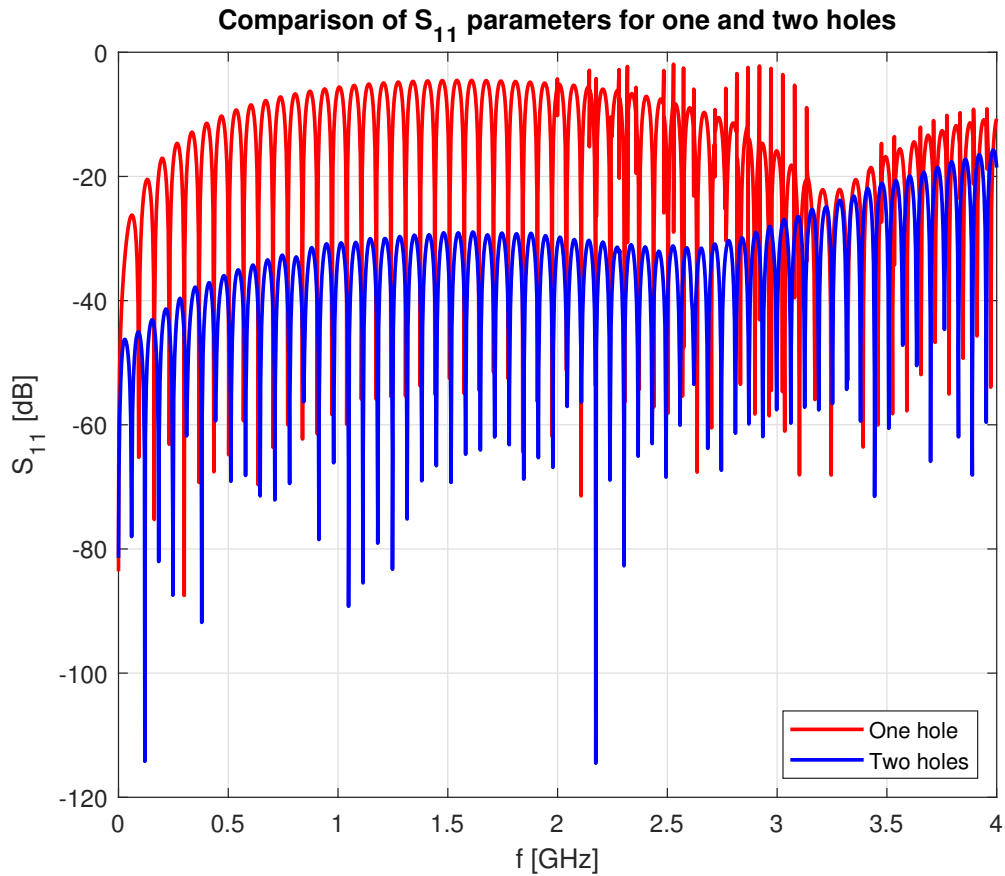


Fig. 6.12: Comparison of S_{11} parameter for the design with 1 and 2 holes.

This simulation was done only for the stripline with no transition to the coaxial line. Then we added the transition from the stripline to the coaxial line. The transition is similar to the previous design and its illustration can be seen on 4.5. The geometry of the strip is changing smoothly to the coaxial line geometry. The supports end a few millimeters before the transition and the width of the strip is

higher at this point. We tuned this transition region to achieve the lowest reflection. After the tuning, we simulated the full-length design with all parts and simplified geometry. For the simulation, we did not include parts such as flanges and screws. They should not affect the simulation results but the simplified geometry allows the simulation to be faster. The results of the simulation can be seen in Fig. 6.13.

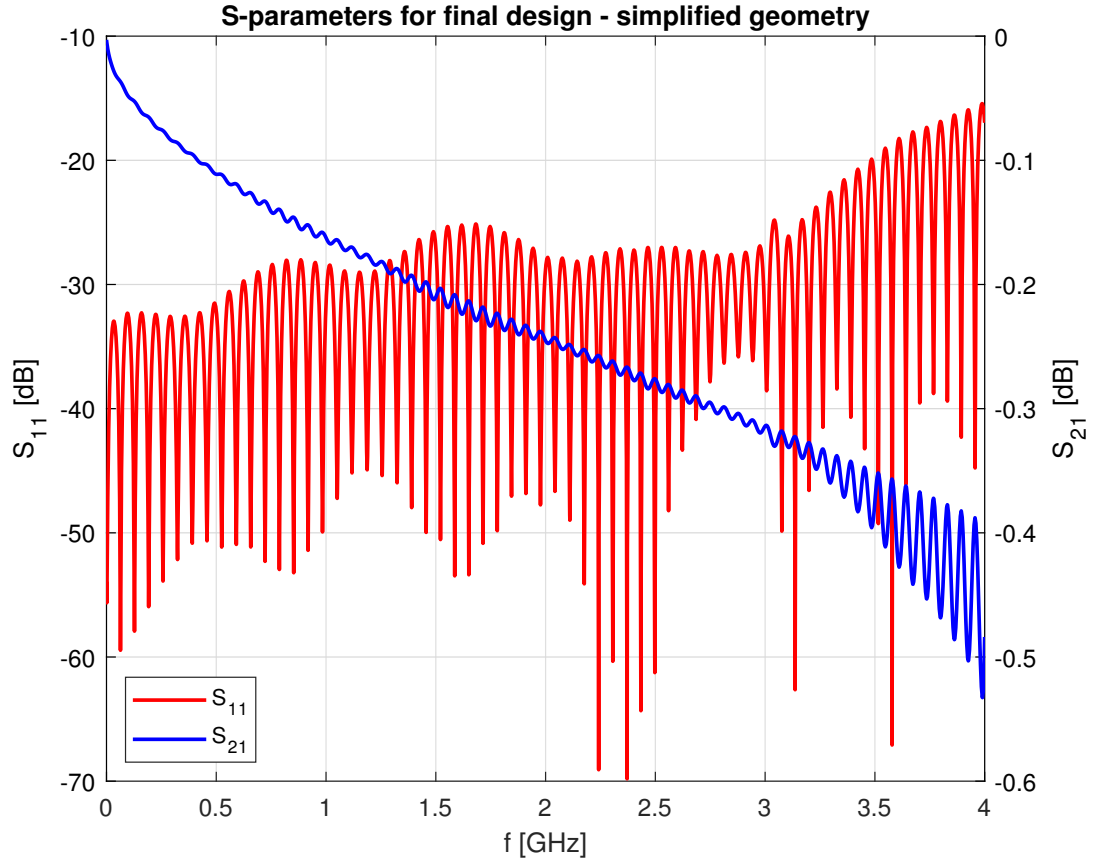


Fig. 6.13: S-parameters for the simplified design.

As can be seen in Fig. 6.13 the reflections are below -20 dB up to 3.5 GHz. That is a significant improvement compared to the previous designs. It was decided to keep this version as the final design.

7 Construction of the final design

We again modeled the whole design in Autodesk Inventor. Technical drawings of all custom parts can be seen in appendix A and we made in Autodesk Inventor as well.

The first step in assembling the stripline is to create the middle part of the outer conductor. It consists of 2 m long aluminum tube and aluminum flanges at the ends. The flanges will be welded to the tube. For the proper welding, the tube needs to be modified at its ends, some material needs to be removed and the edges at the end need to be chamfered.

The flanges are custom and will be CNC machined. For proper welding, the edge that is closest to the tube is also chamfered. In the flange, there is also a groove. The groove is also necessary for proper welding. The connection of the tube and the flange can be seen in Fig. 7.1.

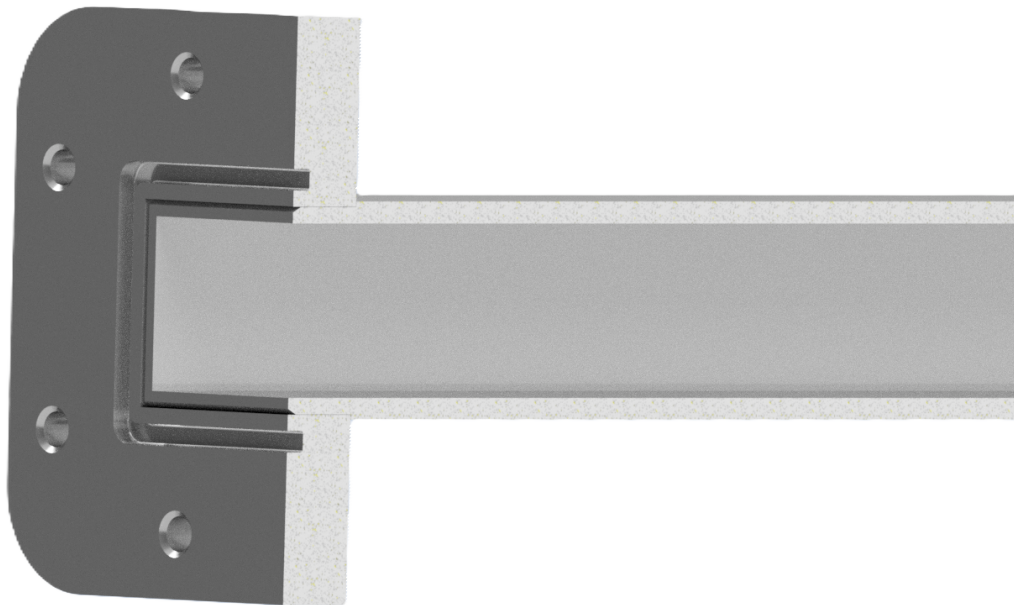


Fig. 7.1: Final design - middle part of the stripline outer conductor.

The center conductor of the stripline will be then inserted with the supports to the tube. The center conductor will be made from a copper rod and modified at its end for transition to the coaxial line. The supports will be CNC machined of teflon. Machining roughly 2 m long part is complicated. Thus the supports will be machined as several shorter pieces and then put there together. The center conductor with the supports inserted into the middle part of the outer conductor can be seen in Fig. 7.2.

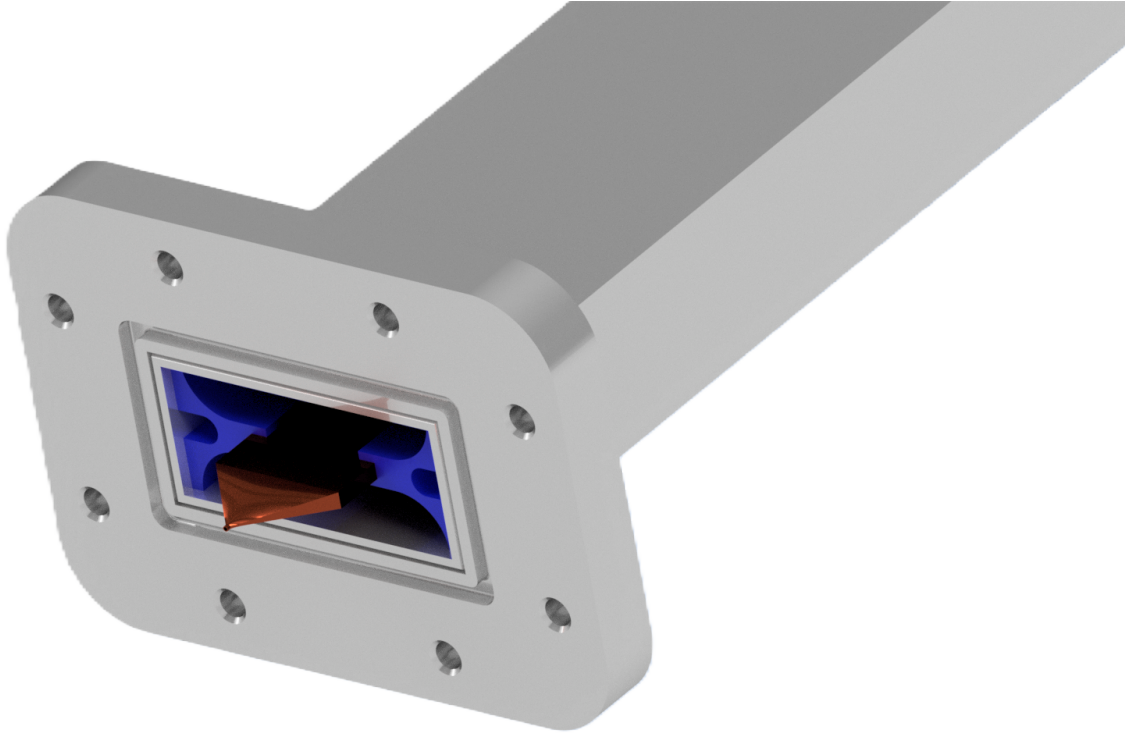


Fig. 7.2: Final design - middle part of the stripline.

End parts of the stripline outer conductor will be CNC machined from aluminum. On its front, there is a flange for connecting to the middle part. In the flange, there is a groove for keeping the design vacuum tight. On the back, there is an ISO K 100 flange for connecting to the rest of the experiment setup. The end part can be seen in Fig. 7.3.

On the side of the outer conductor, there is a double-ended SMA feedthrough with ISO KF 16 bulkhead clamp BCH16. Inside of this apparatus will be a coaxial line with one end connected directly to the center conductor of the stripline and on the other end will be SMA connector connected to the feedthrough. A cut through this part from the top can be seen in Fig. 7.4.

The middle section of the front is chamfered. In the middle, there is a hole for the coaxial line. Through this 3.6 mm hole will be dragged coaxial cable. There are also two bigger holes with diameter 14 mm. These holes serve as entry or exit for the beam and also for pumping out the air from the middle part of the apparatus. A cut through this part from the side can be seen in Fig. 7.5.

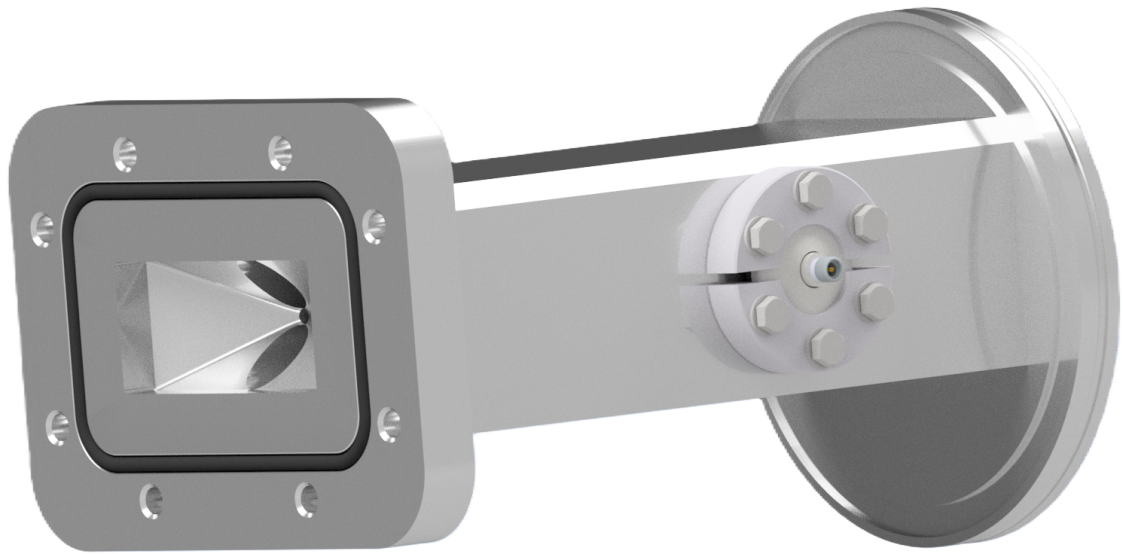


Fig. 7.3: Final design - end part of the outer conductor.

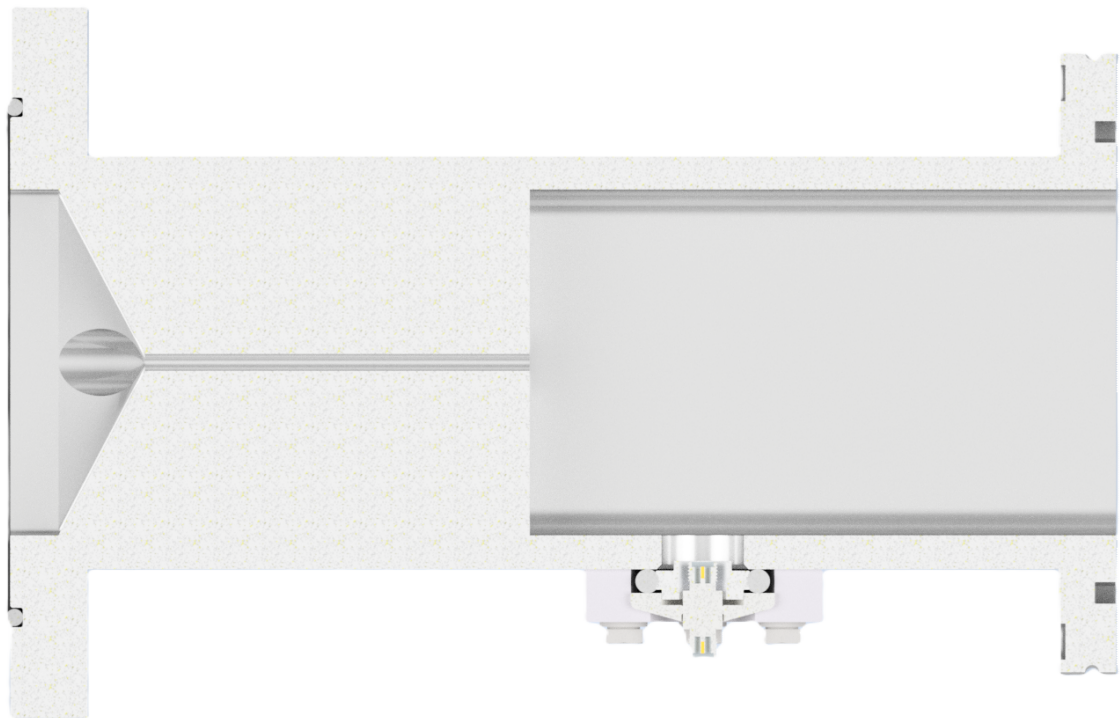


Fig. 7.4: Final design - end part, cut from top.

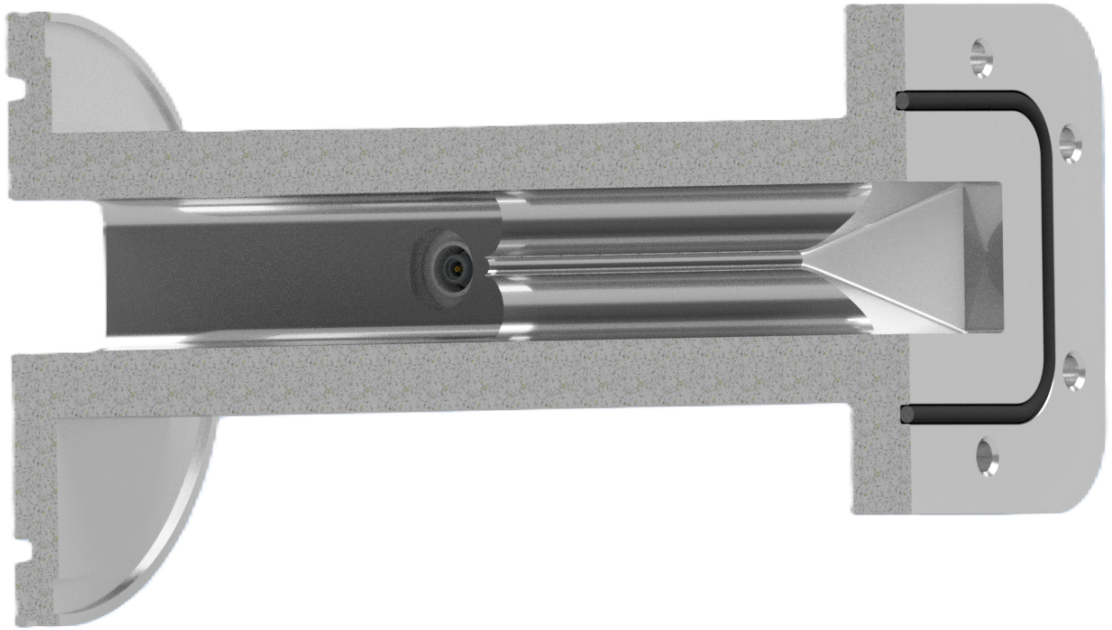


Fig. 7.5: Final design - end part, cut from side.

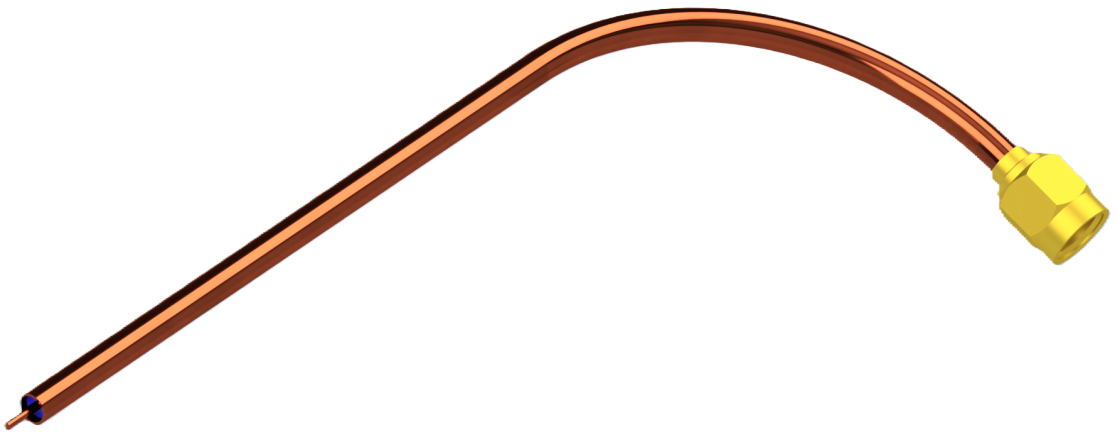


Fig. 7.6: Final design - coaxial line.

The coaxial line will be created of semi-rigid cable with teflon dielectric RG402 with SMA connector 1050524-1 from TE CONNECTIVITY. At the other end of the coaxial line, the inner conductor will be sticking out and will be inserted into a hole in the center conductor of the stripline. Also 1 mm of dielectric will be removed. After connection to the stripline, this will create a transition from stripline to 1 mm long coaxial line with vacuum as dielectric and then to the coaxial line with teflon as dielectric. The coaxial line can be seen in the following Fig. 7.6 and the connection can be seen later in Fig. 7.7.

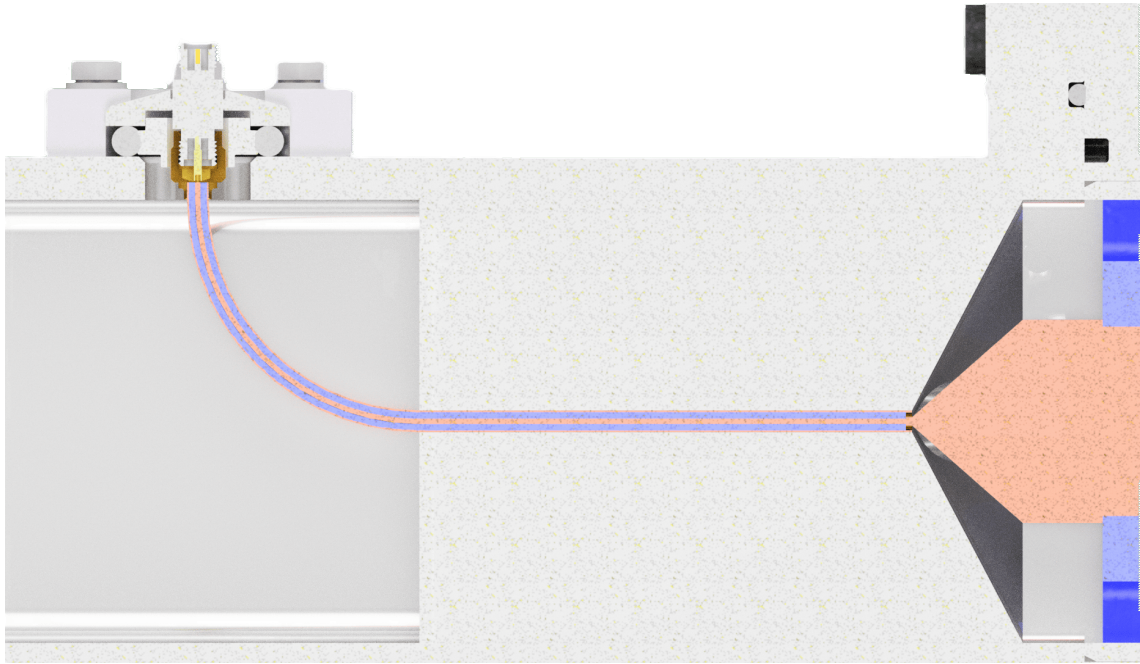


Fig. 7.7: Final design - stripline to coaxial line connection detail.

8 Development of detection electronics for β detector setup for the VITO beamline

The second part of the thesis is focused on the development of detection electronics for β detector setup for the VITO beamline. We designed two different detection PCBs (Printed Circuit Boards) for a small detector and a big detector. Some preliminary images of the detectors rendered by Daniel Paulitsch can be seen in an appendix in figures C.2 C.3 C.4. The difference between these two boards is only dimensions and shape adjustments. The responsible designer of the detector is Daniel Paulitsch. This project serves as his PhD thesis in ISOLDE facility at CERN. Except for the detector PCBs we designed also summing PCB that sums signal from all of the detector PCBs, power distribution PCB for powering all of the PCBs and optional filter PCB.

8.1 β -detector

β detected NMR can be used to determine the magnetic moments of short-lived isotopes with higher sensitivity than conventional NMR. The next isotope to be investigated with β detected NMR at VITO is ^{11}Be . For such measurements the β -detectors are crucial. Their purpose is to detect asymmetric emission of β -particles at 0 and 180 degrees to a magnetic field direction from the hyperpolarized decaying isotopes. The β -detection setup consists of two detectors – the big detector and the small detector. [30, 31]

β -NMR is type of NMR that is based on β decay of a polarized ensemble. It extends the capabilities of conventional NMR to study of magnetic properties of radioactive nuclei. The NMR resonances can be observed as changes in the β decay asymmetry which significantly increases the signal strength. β -NMR relies on a beam of hyperpolarized short-lived β -decaying nuclei. [30, 31]

8.1.1 β particle detection

The detection of β particles starts with a scintillator. The scintillator is a material that emits (fluorescence) photons when a β particle passes through it because the latter excites the electrons in the atoms of the material. [32]

For the photon detection, we used SiPMs Silicon PhotoMultipliers. SiPM is a component similar to a photodiode. It is a photodetector that generates short current pulses with a very steep rising edge as a response to photon absorption.

SiPMs are highly sensitive with very low time jitter that allows them to capture single photons. [33, 34].

Between the scintillator and the SiPMs, we introduced a plastic diffuser that allows a similar fraction of emitted photons from the scintillator to reach the SiPMs, independent of the position of the β particle interaction. In our design, we are using AFBR-S4N66P024M SiPM. This SiPM was selected by Daniel Paulitsch according to its good parameters and low price. The image of the SiPM can be seen in the following Fig. 8.1. [35]

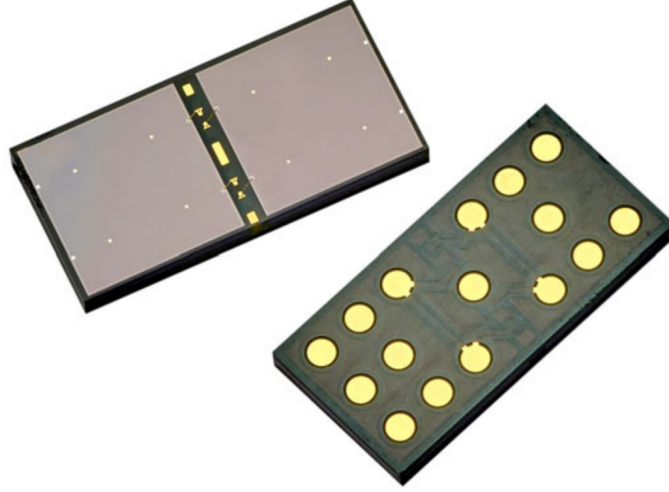


Fig. 8.1: SiPM selected for the detector. [35]

8.2 The experiment setup

The experimental setup for β detection part of β -NMR consists of two β detectors. They consist of a plastic material that emits visible-light photons under impact from β particles and several SiPMs. The last part of the detection setup is a DAQ (Data Acquisition System) used for processing the signals from the detectors. The block diagram of the setup can be seen in Fig. 8.2.

The DAQ consists, among others, of an attenuator/amplifier, oscilloscope, and host PC (Personal Computer) for controlling the DAQ. The oscilloscope that is used is NI PXIe-5170R. Its input voltage has fixed range of -5 V to 5 V . This needs to be considered in the design so as not to overshoot this voltage limit. Optionally the signals can be filtered before entering the DAQ.

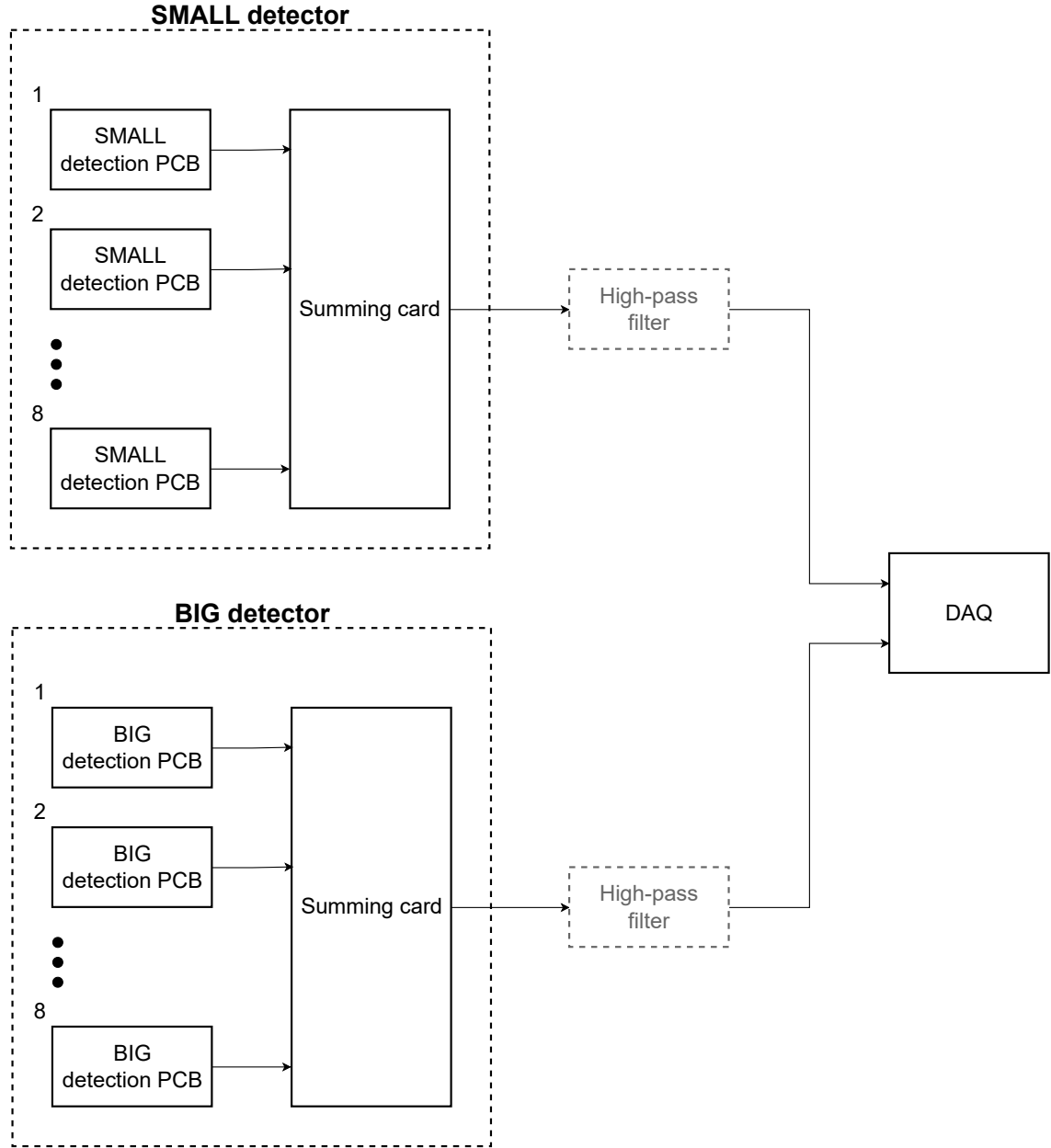


Fig. 8.2: The experiment setup.

8.3 Detector PCB design and simulation

The new β -NMR measuring system should be able to also measure the energy of the β particles. The energy of the particles is directly proportional to the number of emitted photons. For measuring the energy, the linearity of the detection PCB is strongly advantageous.

The whole design is also placed in a strong magnetic field of 4.7 T that is necessary for the β -NMR measurement. The design needs to be magnetic field compatible thus

all components need to be non-magnetic or very weakly magnetic, not to be affected by the magnetic field and also not to disturb the magnetic field. This is especially crucial for big metallic parts, such as connectors.

The whole detector system will be placed in a high vacuum that is also necessary for the β -NMR measurement. Thus, the whole design needs to be vacuum-compatible. This is important mostly for plastic parts of the design which can reduce the vacuum quality because of outgassing (when a small amount of gas escapes from the plastic material). This is especially crucial for big plastic parts such as cables.

The simplified connection of detector electronics can be seen in Fig. 8.3. The signals from SiPMs generated by photon absorption are added together and amplified. The signal from the amplifier goes to SMA connector and is read out.

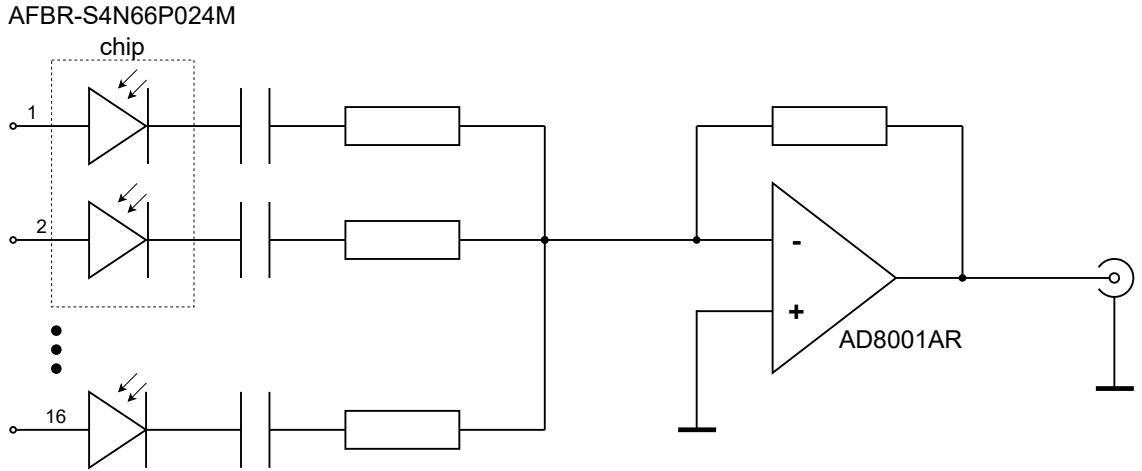


Fig. 8.3: Simplified circuit.

The design of the detection PCBs was inspired by a board designed at DPNC (Department of Nuclear and Particle Physics) at the University of Geneva. The selected amplifier *AD8001* [36] is the same as in the University of Geneva design. They are using the amplifier for a similar purpose thus we used the same one and saved some time not searching for another. We made a few adjustments to the design to achieve desired functions such as for example linearity and amplification.

One of the key aspects of the design is linearity. We succeeded in designing such a circuit and we simulated it in LTspice with SiPM model provided to us by the University of Geneva. We just adjusted the model for our specific SiPM *AFBR-S4N66P024M*. We managed to obtain the desired parameters. The dependency of output voltage on a number of detected photons and its dependency on a number of cells can be seen in Fig. 8.4 for 30, 60, 120, 200, 400 and 500 detected photons. The gain of the amplifier for this simulation was set to 2 and the SiPM's overvoltage was set to 12 V.

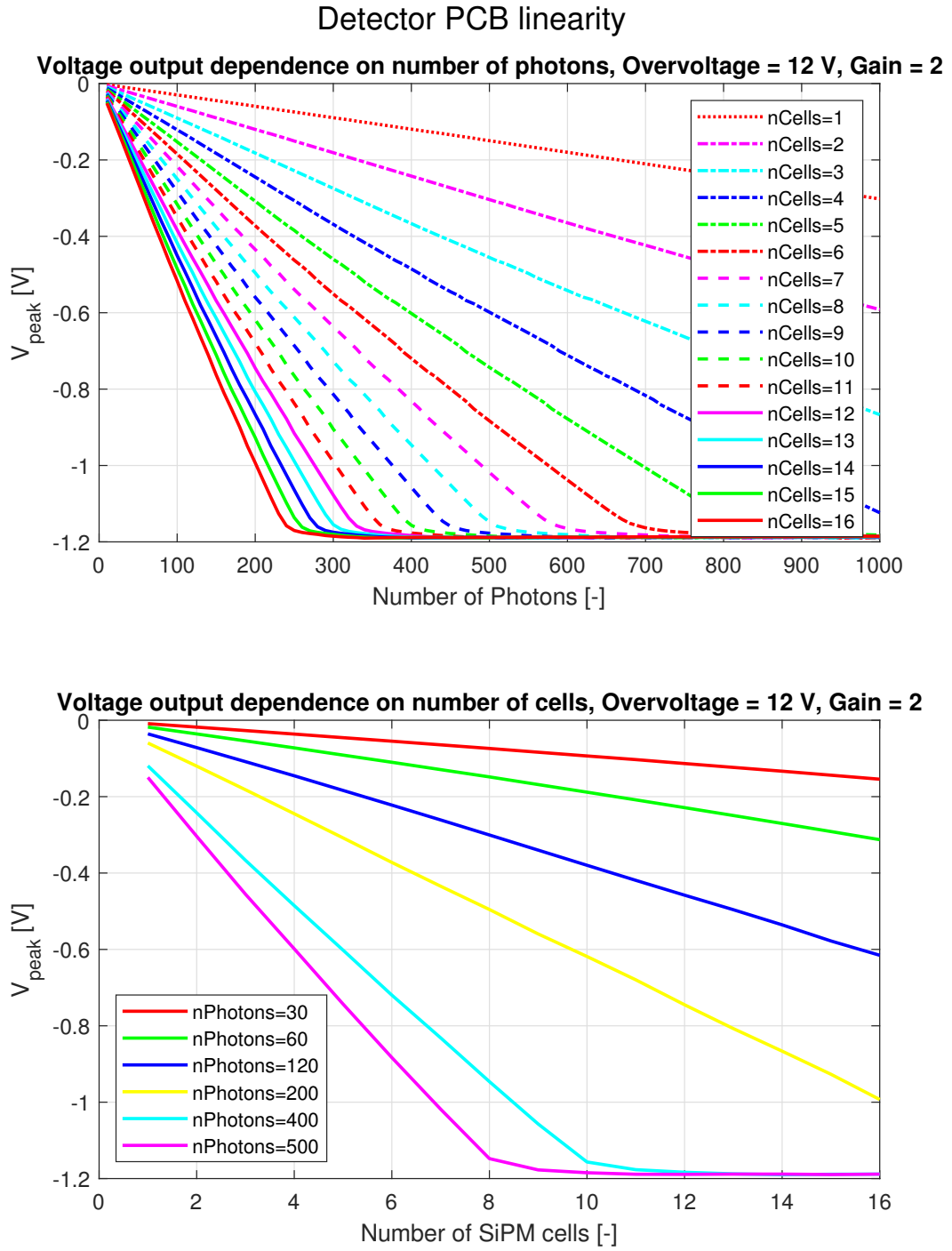


Fig. 8.4: Output voltage dependent on number of detected photons and number of cells.

Initially, we thought of using 4 SiPMs which each has 2 cells. Thus 8 cells per one PCB and 16 PCBs. But later we decided to use 8 SiPMs per board and 8 PCBs instead due to space concerns. The full schematic of the detector electronics can be seen in appendix B.1.1.

As we can see in Fig. 8.4, the design is linear, up to roughly 200 photons absorbed by all sixteen SiPMs. This range is sufficient for the first planned isotope ^{11}Be . The expected maximum number of photons for ^{11}Be beta particles is 116, according to the results of simulation (using GEANT4 software) performed by Daniel Paulitsch. At some point, the circuit stops being linear. It is caused by the saturation of the amplifier. When the SiPMs detect too many photons, their output signal becomes too large and the amplifier is saturated. But this can be easily solved by changing the gain of the amplifier.

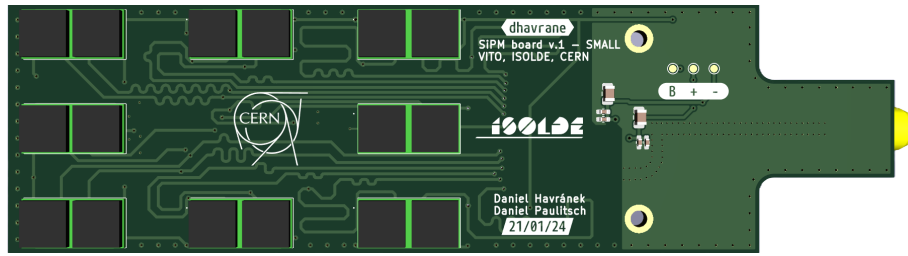
8.4 Detector PCB design

The small and big detector PCB needs to have specific dimensions to fit inside of the detector. For better detection, SiPMs had to be placed at the corners of the PCB and no components between them on the same side of the PCB. For the *Big* PCB this means that all components need to be on the other side of the PCB. The *Small* PCB is extended so the connectors can be reachable. Thus, some small components are allowed on the other side. This also limits the design to SMD (Surface Mount Device) components only.

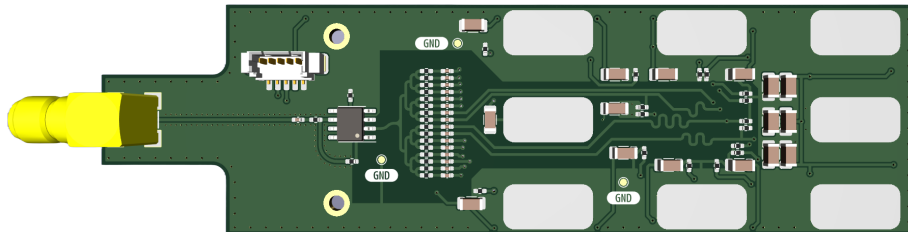
Another important aspect of the PCB design is the length matching of the signals. Each trace needs to be the same length, otherwise, the signal from each SiPM arrives at a different time, causing the resulting signal to be distorted.

Optionally the PCBs can be cooled to low temperatures (down to -40°C) to lower the noise. The PCBs need to be designed with the option of cooling them. The temperature needs to be the same for all SiPMs to achieve low noise and good stability. For this, we added cooling pads on the other side of the PCB below the SiPMs. For the cooling, a custom aluminum heatsink will be used with a commercial chiller. The PCB design of the *Small* PCB can be seen in figures 8.5 and 8.6.

The SMA connector on the *Big* PCB that can be seen in Fig. 8.6 is heading towards the center of the PCB. This is done on purpose because of the custom aluminum heatsink. The heatsink will be all around the edges of the PCB as can be seen in an appendix in Fig. C.2 of the *Big* detector design. This way it leaves space for easier cable attachment.

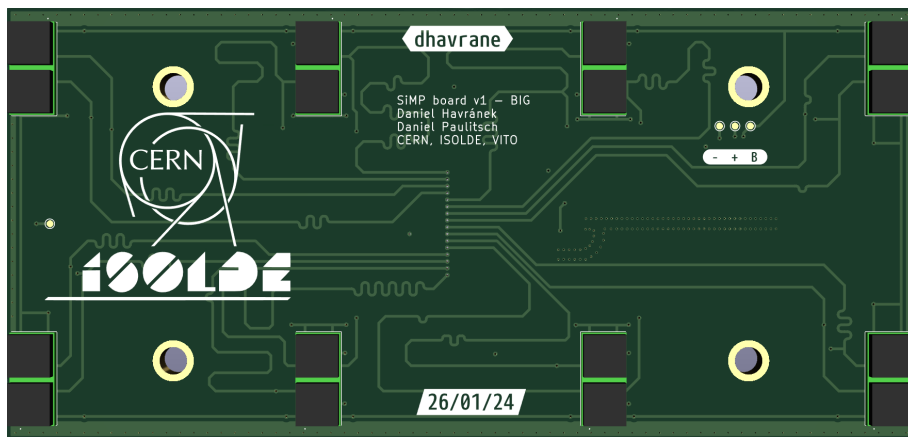


(a) Top

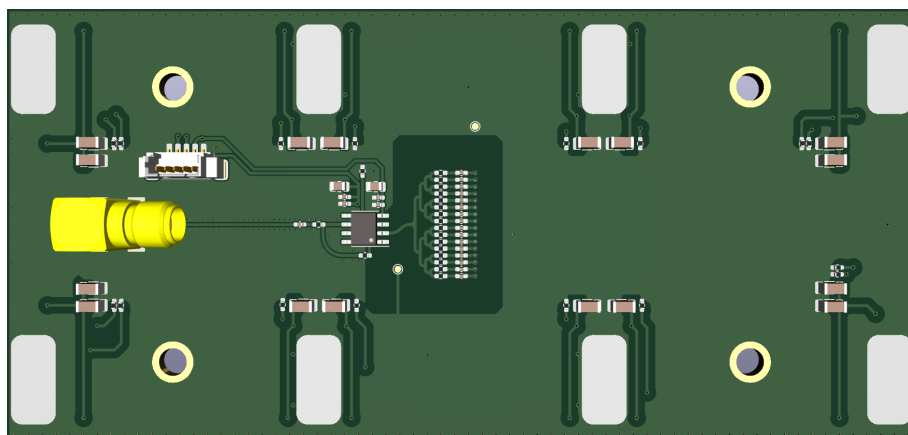


(b) Bottom

Fig. 8.5: Detector PCB - *Small*.



(a) Top



(b) Bottom

Fig. 8.6: Detector PCB - *Big*.

8.5 Other PCBs

Among the other PCBs belong the *Summing card*, *Power distribution* PCB and optionally the filter PCB. The *Summing card* and the *Power distribution* PCB may or may not be in the magnetic field. It is dependent on the detector construction which is not yet finalized. Thus PCBs should be magnetic field compatible and vacuum compatible too. The filter PCB will be outside of the vacuum and magnetic field.

8.5.1 Summing card

The summing card combines signals from 8 detection PCBs to one output. All 8 inputs go again to a summing amplifier with a specific gain not to overshoot the oscilloscope's range. The layout of the summing PCB is similar to that of the *Small* and *Big* detector PCBs. The image of the summing card can be seen in Fig. 8.7.

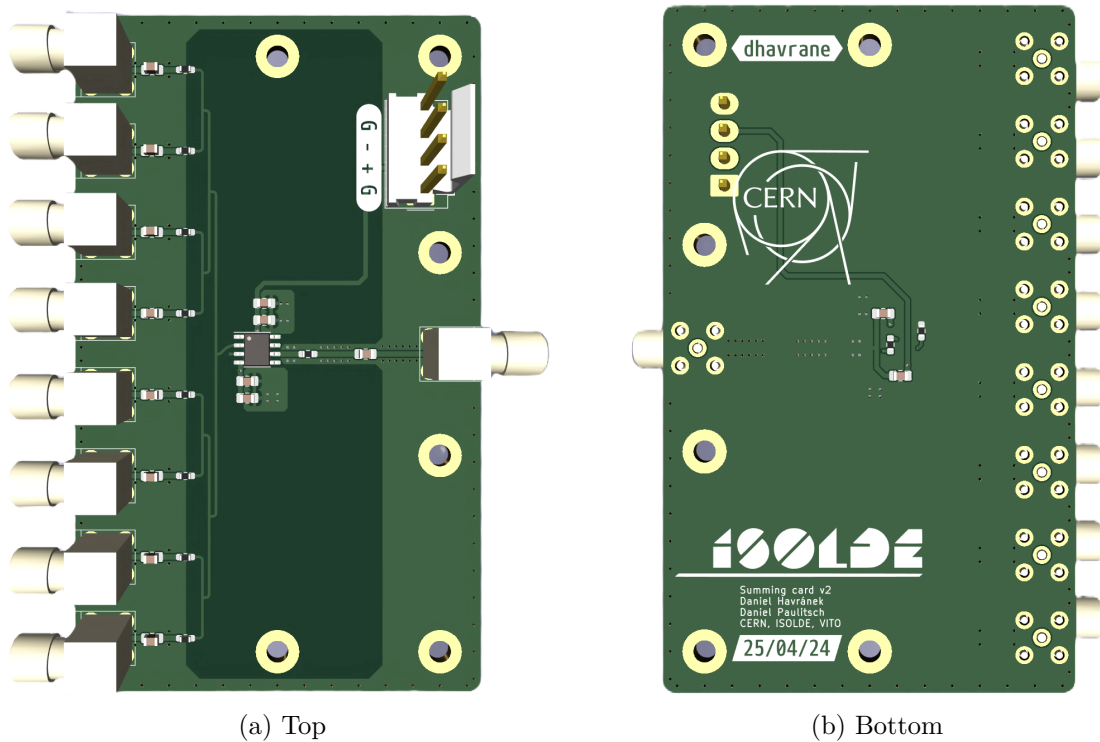


Fig. 8.7: Summing card.

Initially, we used the same amplifier AD8001AR. But this amplifier needs to be changed. The signal from all of the 8 detector PCBs will be probably too high and the amplifier will be saturated.

As a new amplifier we chose AD8007ARZ [37]. It is also a low-noise amplifier. Its bandwidth and slew rate are a bit smaller but sufficient. The whole schematic of the summing card can be seen in appendix B.1.2.

We simulated the circuit of the summing card in LTspice. The output pulse can be seen in Fig. 8.8. The output pulse was simulated for 8 detection PCBs. From the simulation of detection PCB the output pulse for 16 SiPMs cells and 116 photons was exported. In the simulation of the summing card, the detection PCBs were replaced by a pulse voltage source and imported the exported pulse from detection PCB.

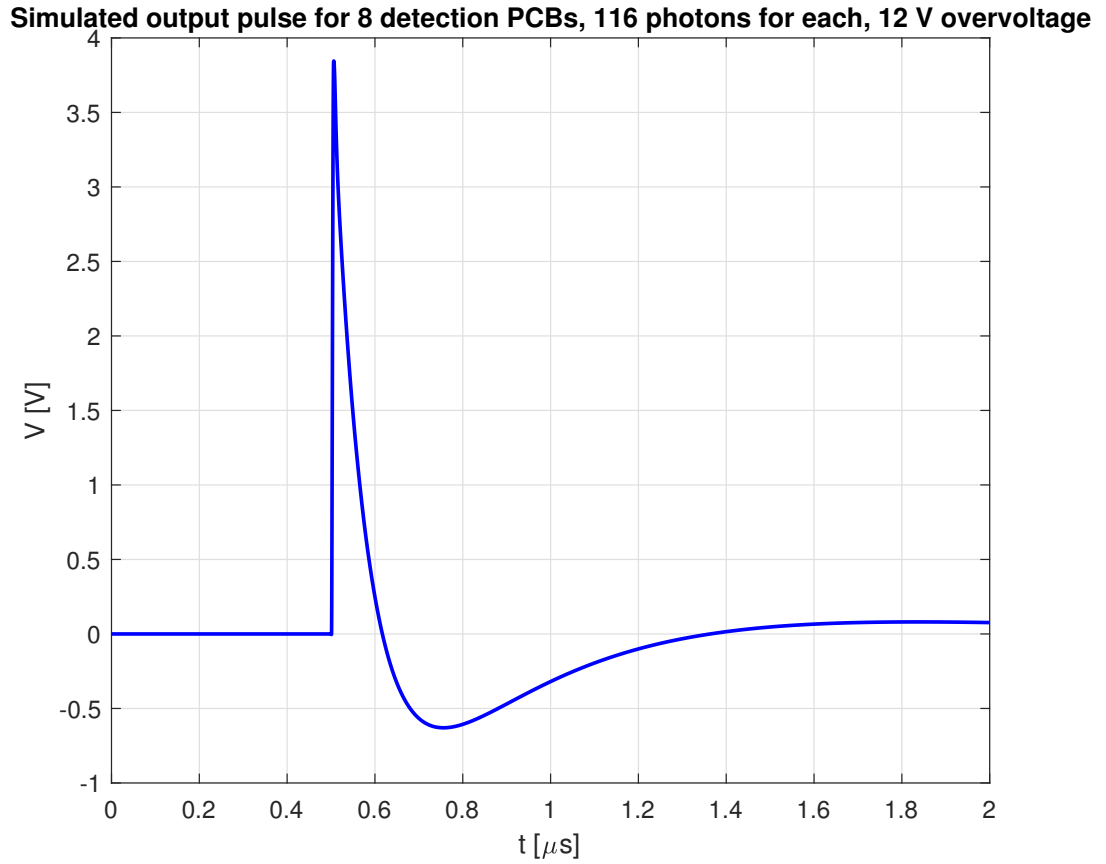


Fig. 8.8: Summing card output pulse for 8 detection PCBs

We also checked the linearity of the summing card in the LTspice simulation. We simulated output voltage peak dependency on a number of input detection PCBs. The results can be seen in Fig. 8.9. As can be seen, the design is almost linear and its properties are sufficient.

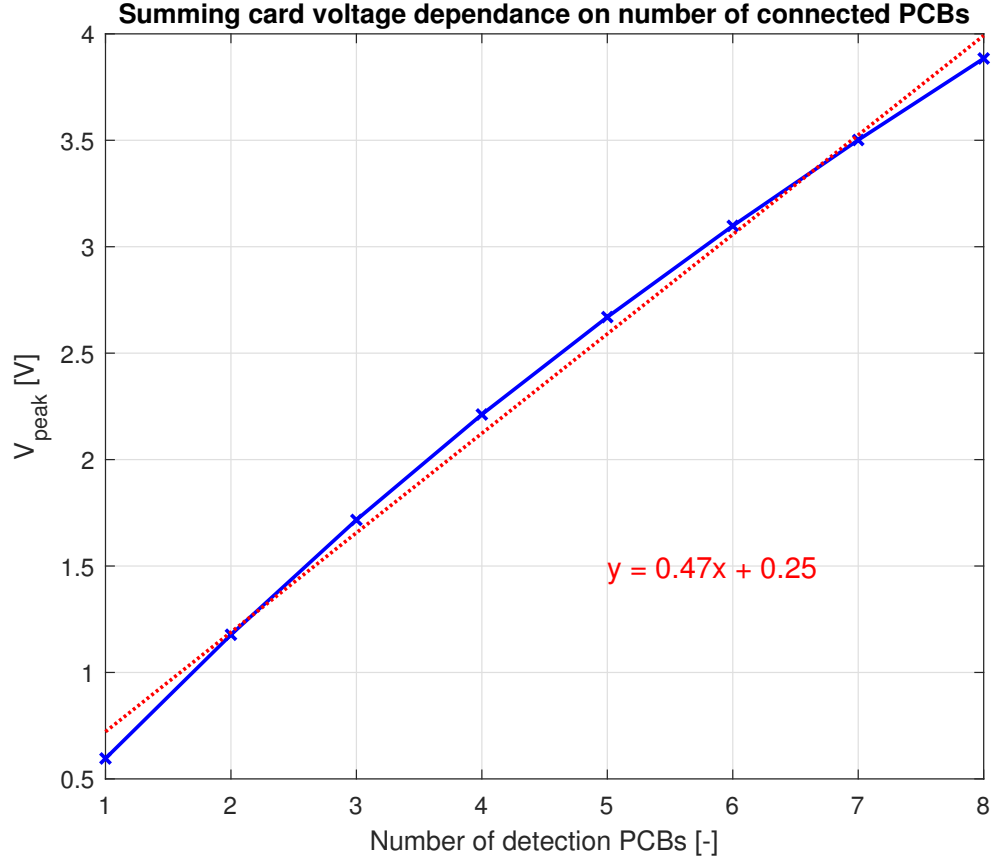


Fig. 8.9: Summing card linearity dependency on number of detection PCBs.

8.5.2 Filter PCB

The filter PCB is optional. It is a simple RC (Resistor-Capacitor) high-pass filter with 47 pF capacitor and 50 Ω resistor. This filter can remove the low-frequency components of the signal resulting in the signal being narrower which can be useful in the future. The filter PCB can be seen in Fig. 8.10.

It cannot be used for the energy measurement. For energy measurement, the whole output signal from the SiPMs is necessary for proper recalculation of the output voltage to the energy. In the case of count rate measurement the shape of the output SiPM signal can be changed. In this case, the filter PCB can be used if necessary.

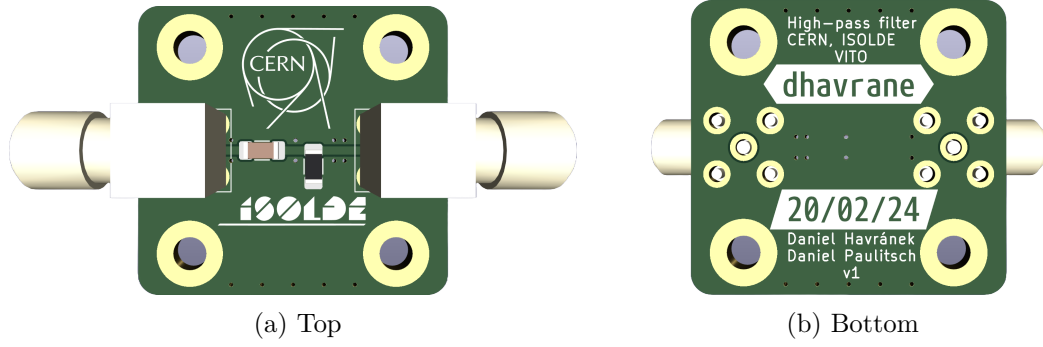


Fig. 8.10: Filter.

8.5.3 Power distribution PCB

The power distribution PCB serves only for power distribution for the detector PCBs. The whole design is sealed in vacuum-tight detector. The number of cables going in and out is limited by the space available for electrical feedthroughs. Due to these reasons, it is better to have a lower number of cables going in and then split them. The Power distribution PCB can be seen in Fig. 8.11. The whole schematic for the *Power distribution* PCB can be seen in appendix B.1.3.

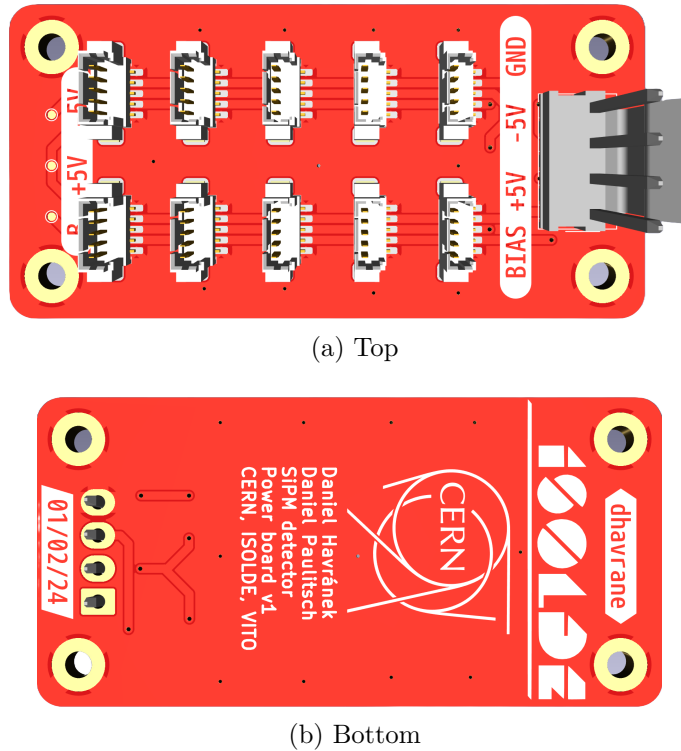
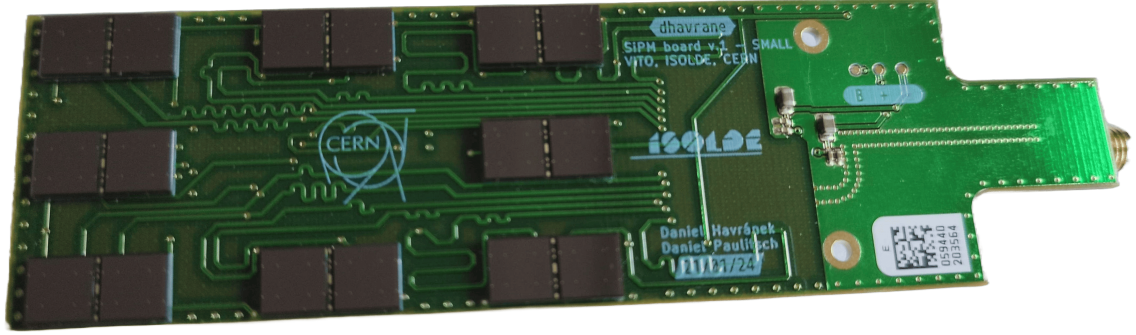


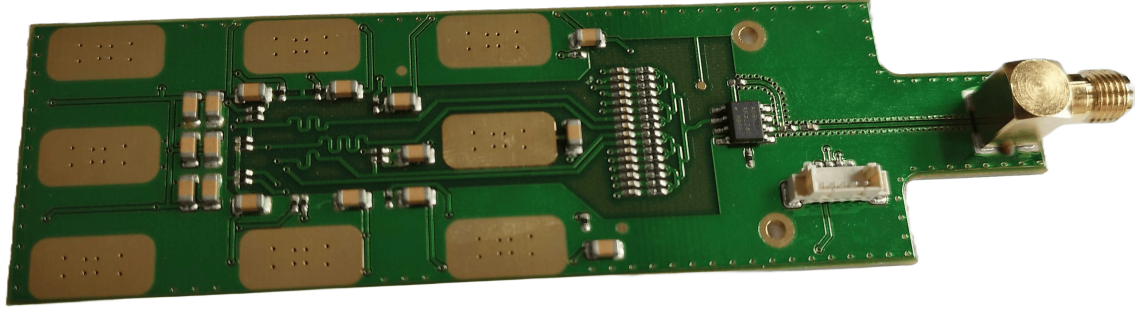
Fig. 8.11: Power distribution PCB.

9 Detector PCB testing

For testing the functionality of the detector PCBs we ordered a prototype version of the *Small* detector PCB. We ordered the PCB and its assembling from a Czech company called Printed. Photo of the PCB can be seen in Fig. 9.1.



(a) Top



(b) Bottom

Fig. 9.1: Photo of the *Small* detector PCB.

For the PCB functionality testing we put a piece of scintillator on top and attached it with teflon tape and black electrical tape. We put the PCB into a light-tight box and we performed multiple measurements.

For measuring we used the previously mentioned DAQ system but without the amplifier/attenuator. We went directly to the NI PXIe-5170R oscilloscope and measured signals from the PCB. For controlling the oscilloscope we used a program written in LabView. The pulse from the PCB was measured and the integral of each pulse was calculated. Then peak values and the integral values together with a timestamp were stored in memory and we used them later for data analysis. The peak values and the integral values are important for getting the number of photons detected by the PCB.

For the PCB testing we used several different radioactive sources. As β sources we used ^{90}Sr and ^{207}Bi . As a γ we used ^{137}Cs and ^{60}Co . We also measured background with no source for comparison and for obtaining more precise results. Mea-

sured peak voltages for these sources can be seen in Fig. 9.2. The measured peak voltages are time-normalized because each source was measured a different amount of time. After the time normalization, the background was subtracted from the measured results of each radioactive source. For the measurement a different trigger level was used for each source thus each source starts at a different voltage in the graph. For comparison, the normalized background can be also seen in Fig. 9.2 plotted with a dashed line.

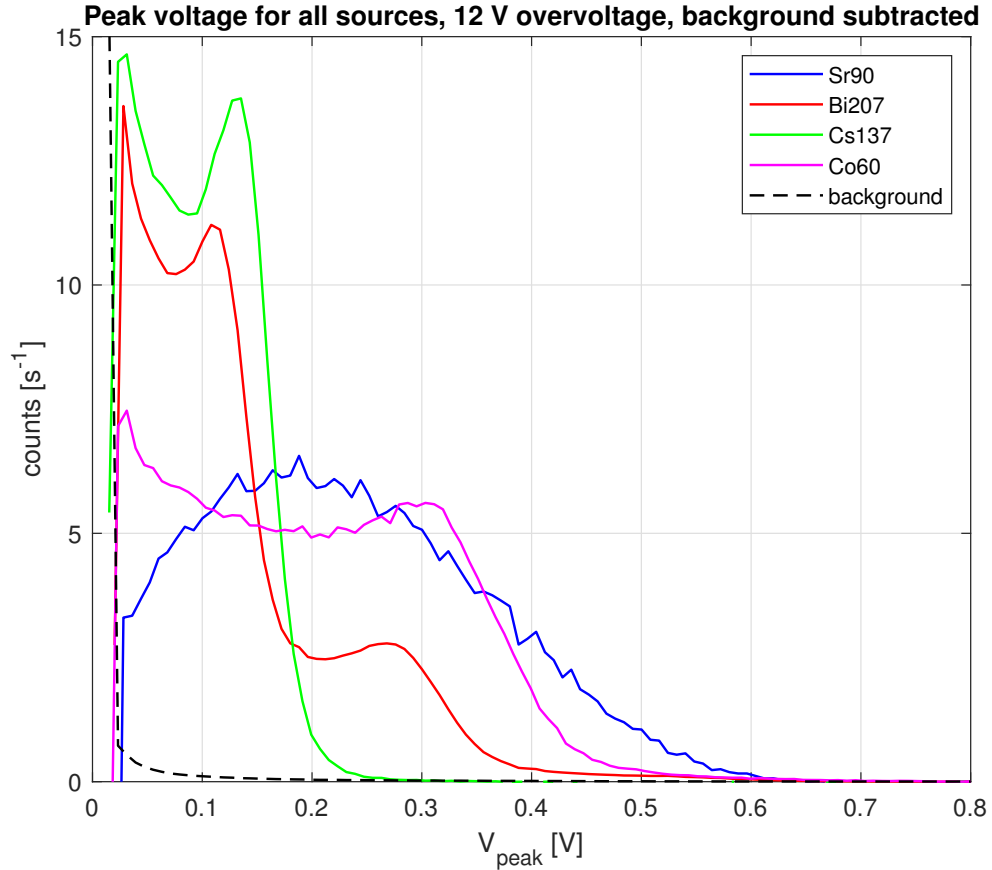


Fig. 9.2: Peak voltages for measured sources with 12 V overvoltage and with background subtracted.

The peak voltage is good for a fast check of the results and their comparison to expected values. For precise measurement of the number of detected photons, the integral value is necessary. If we integrate the voltage pulse by time and divide by amplification of the PCB amplifier and SiPM amplification (dependent on overvoltage) we receive values with unit V s. If we divide it with specific line impedance which is $50\,\Omega$ we receive values with unit A s what is electric charge in C. If we divide the values now with an absolute value of the charge of an electron (which is

$-1.602 \cdot 10^{-19} \text{ C})$ we receive the number of detected photons. The whole equation for the calculation of the number of detected photons is

$$N_{photons} = \frac{1}{|e| Z G_{amp} G_{SiPM}} \int_{t_1}^{t_2} v(t) dt, \quad (9.1)$$

where $N_{photons}$ is the number of detected photons, e is charge of an electron, Z is line impedance, G_{amp} is gain of the amplifier, G_{SiPM} is gain of the SiPMs, t_1 is the beginning of the pulse, t_2 is the end of the pulse and $v(t)$ is the voltage which is function of time.

The number of detected photons for each source can be seen in Fig. 9.3. The measured data are again time normalized, the background is subtracted and the measurement was done for 12 V overvoltage and with different trigger levels for each source as in the previous case.

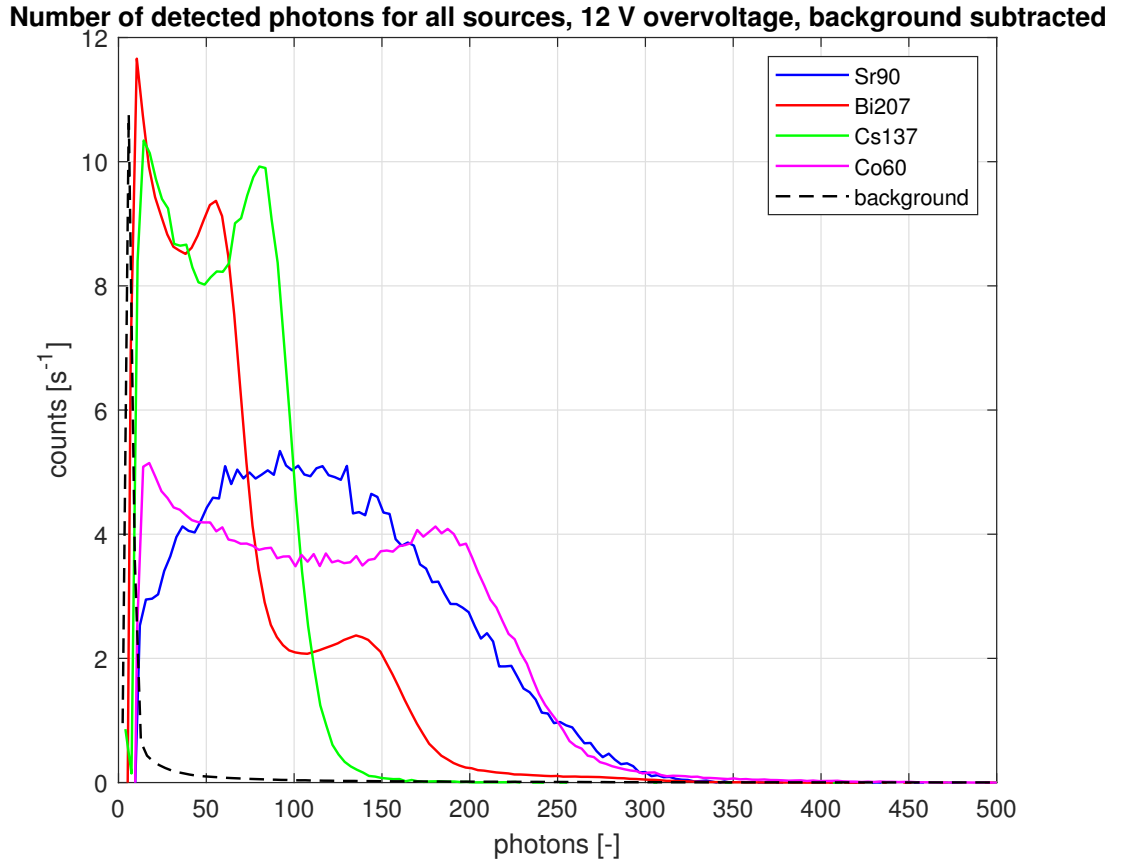


Fig. 9.3: Number of detected photons for measured sources with 12 V overvoltage and with background subtracted.

We also measured some of the sources for different overvoltages. The results of these measurements can be seen in appendix B.3. To be sure we also performed

a linearity check on overvoltage. The measured results can be seen in Fig. 9.4. As can be seen, the PCB has linear behavior. Small deviations can be caused by eyeballing the results. At 13.5 V overvoltage the PCB's amplifier is saturated for used source ^{90}Sr . We performed this measurement with DAQ system with the use of its additional amplifier.

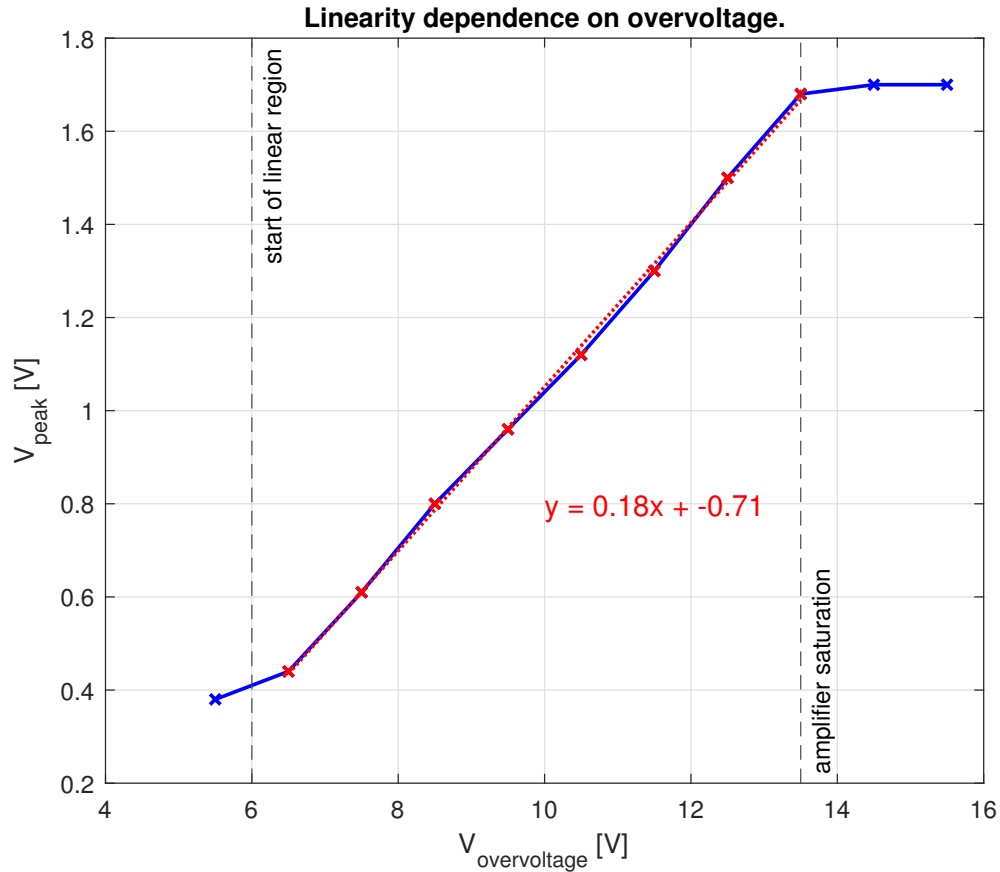


Fig. 9.4: PCB linearity dependency on overvoltage - measured with ^{90}Sr and with DAQ amplifier.

Conclusion

We have designed and simulated an initial design for RF-atom/ion beam interaction region. The design was focused on developing a transmission line with a uniform RF field. Therefore a vacuum spaced stripline geometry was employed. Another goal was to minimize reflections thus multiple ways of stripline to coaxial line connections were simulated.

According to the simulation results, we have selected the best design for the considered frequency range from 0 Hz up to 1 GHz and we have modeled it in CAD software. We have created technical drawings for the design and the design was ready for manufacturing.

The initial design was supposed to be used in the November 2023 beamtime. Unfortunately, an accident happened during the experiment preparation and the November beamtime was cancelled. The experiment was moved to fall 2024 so the initial design was not needed anymore and it was decided, that the initial design would not be manufactured.

We redesigned the apparatus to obtain a higher frequency range. The final design is usable in significantly higher frequencies up to roughly 3.5 GHz and fulfills expectations. After the final design is manufactured it will be measured. After the measurement, the apparatus will be used in the experiment in the next beamline.

For the β detector the electronics were designed. The detection PCB fulfills the expectation. Linearity of the PCB was simulated and tested with measurement with different β sources.

Magnetic field compatibility was verified with a strong magnet. Vacuum compatibility was achieved by using proper materials and reducing the amount of plastics inside the vacuum chamber. For good cooling, there are cooling pads on the PCB bellow the SiPM detection elements. SiPMs are appropriately spaced for proper detection function. We length-tuned all traces with fast signals minimize the time distortion. Now the electronics are ready and waiting to be tested with the whole detector.

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Abbreviations

BNC	Bayonet Neill Concelman
CAD	Computer Aided Design
CERN	Conseil Européen pour la Recherche Nucléaire
CNC	Computer Numerical Control
DAQ	Data Acquisition System
DC	Direct Current
DPNC	Department of Nuclear and Particle Physics
HMIC	Hybrid Microwave Integrated Circuit
IEEE	Institute of Electrical and Electronics Engineers
ISOLDE	Isotope mass Separator On-Line facility
MIC	Microwave Integrated Circuit
NMR	Nuclear Magnetic Resonance
PC	Personal Computer
PCB	Printed Circuit Board
PEC	Perfect Electronic Conductor
RC	Resistor-Capacitor
RF	Radio Frequency
SIW	Substrate Integrated Waveguide
SMA	SubMiniature version A
SMD	Surface Mount Device
SNR	Signal to Noise Ratio
SiPM	Silicon PhotoMultiplier
SWR	Standing Wave Ratio
TE	Transverse Electric

TEM	Transverse ElectroMagnetic
TM	Transverse Magnetic
VITO	Versatile Ion polarisation Technique Online

Symbols

α_{TEM}	loss of coaxial line
Δ	frequency detuning
Γ_L	reflection coefficient of a load
Γ_S	reflection coefficient of a source
μ	permeability
ω	frequency
ω_0	resonance frequency
Ω_R	Rabi frequency
σ_v	conductance
Θ	simplification coefficient for stripline calculation
ε	permittivity
ε_r	relative permittivity
C_f	fringing capacitance at the edges of the stripline
D	simplification coefficient for stripline calculation
E_{max}	maximal electric field intensity between two conductors
H	distance between center conductor and ground plane of stripline
h	height of stripline substrat
j	complex unit
$P(\Delta, t)$	atom or ion transition probability
P_0^{TEM}	maximal real power of coaxial line propagated by TEM wave
R	resistance
R_0	radius of outer coaxial conductor
r_0	radius of center coaxial conductor
R_S	radius of outer coaxial conductor

SWR	standing wave ration
t	interaction time
t_S	thickness of center conductor of stripline
w	width of center conductor of stripline
w_{ef}	effective width of striplin center conductor
X	reactance
Z	impedance
Z_0	stripline impedance
Z_C	coaxial line impedance
Z_L	impedance of a load
Z_S	impedance of a source

List of appendices

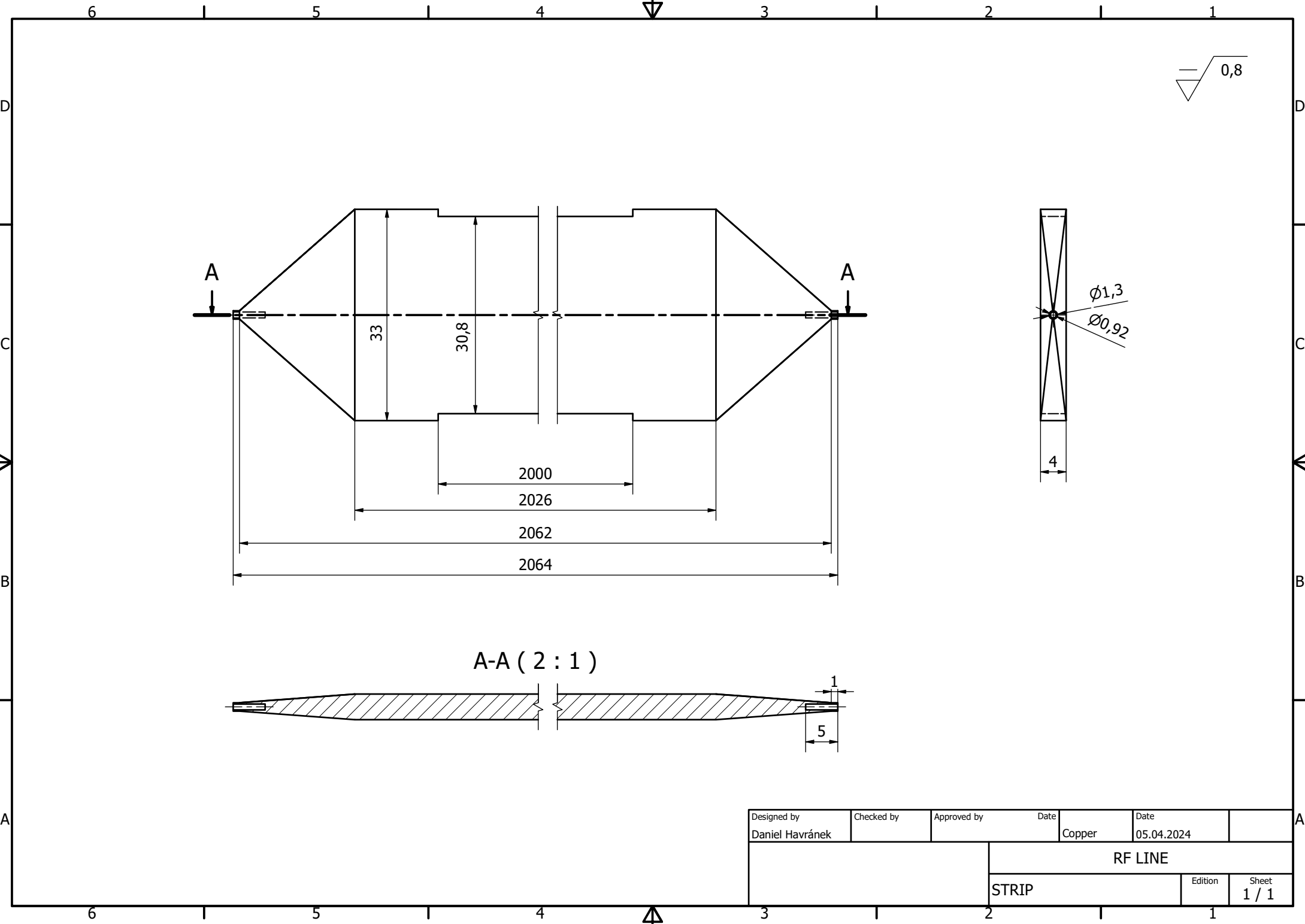
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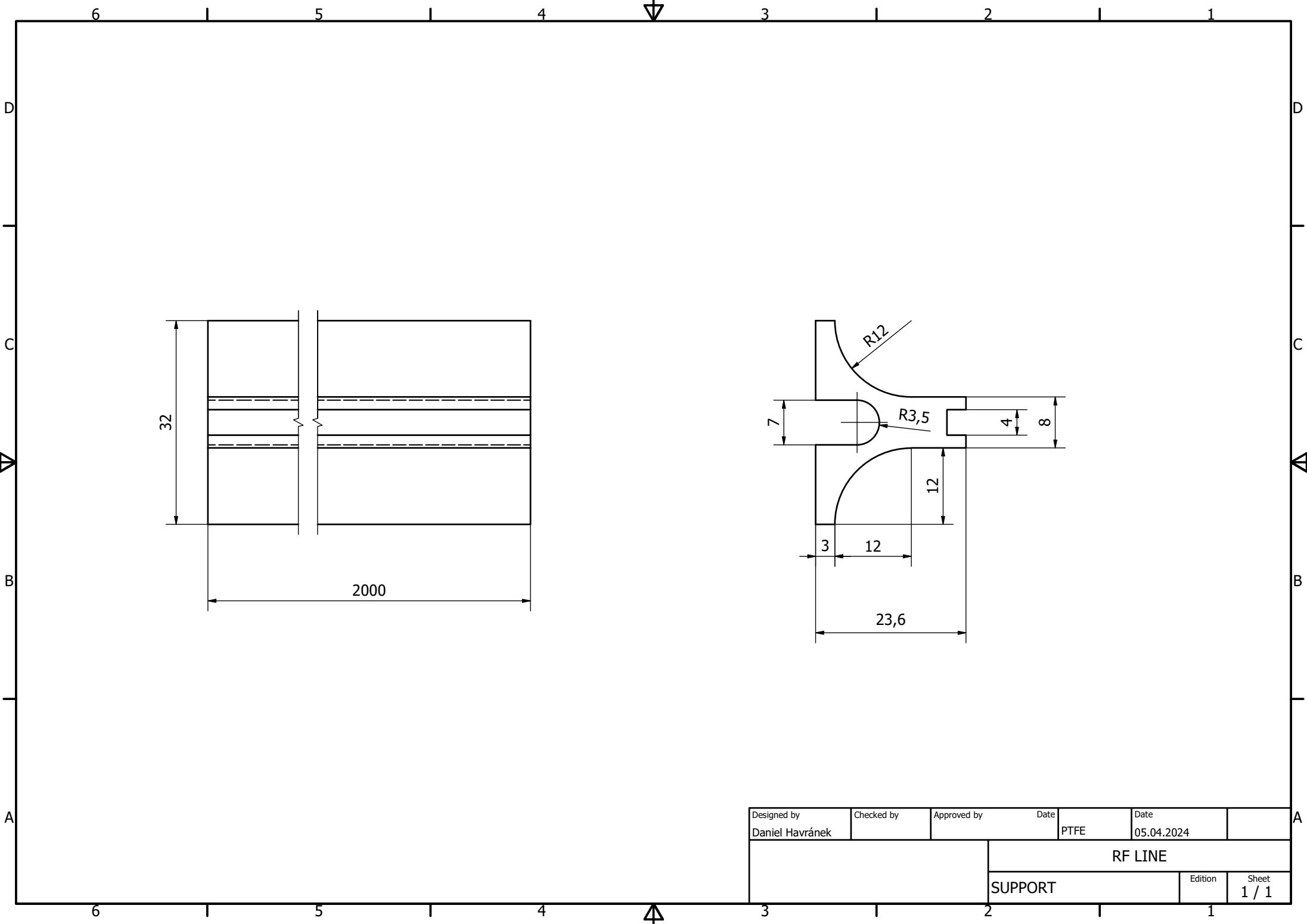
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│   │   ├── BIG.....Big detection PCB
│   │   │   ├── man ..... Manufacturing data
│   │   │   └── SiPM_BIG.kicad_pro ..... PCB project
│   │   ├── FILT.....Filter PCB
│   │   │   ├── gerber.....Manufacturing data
│   │   │   └── FILT.kicad_pro.....PCB project
│   │   ├── lib.....Components library
│   │   │   ├── footprints_symbols.....Footprints and symbols library
│   │   │   ├── logos ..... Logos library
│   │   │   └── step.....3D models of components library
│   │   ├── PWRD ..... Power distribution PCB
│   │   │   ├── gerber.....Manufacturing data
│   │   │   └── PWRD.kicad_pro.....PCB project
│   │   ├── SMALL ..... Small detection PCB
│   │   │   ├── man ..... Manufacturing data
│   │   │   └── SiPM_SMALL.kicad_pro ..... PCB project
│   │   ├── SUMM.....Summing card
│   │   │   ├── gerber.....Manufacturing data
│   │   │   └── SUMM.kicad_pro.....PCB project
│   └── RF_line.....3D model, technical drawings and simulation files of RF line
│       ├── Final..... Final RF line design
│       │   ├── CAD ..... 3D models and technical drawings
│       │   │   ├── lib.....Library with commercial components
│       │   │   └── ASSEMBLY STRIPLINE.iam.....Whole design assembly
│       │   ├── CST.....Simulation files
│       │   │   ├── Final_design_bend.cst ..... Bend version simulation file
│       │   │   └── Final_design_straight.cst ..... Straight version simulation file
│       └── Initial.....Initial RF line design
│           ├── CAD ..... 3D models and technical drawings
│           │   ├── ASSEMBLY STRIPLINE.iam.....Whole design assembly
│           └── CST.....Simulation files
│               └── Initial_design.cst.....Simulation file

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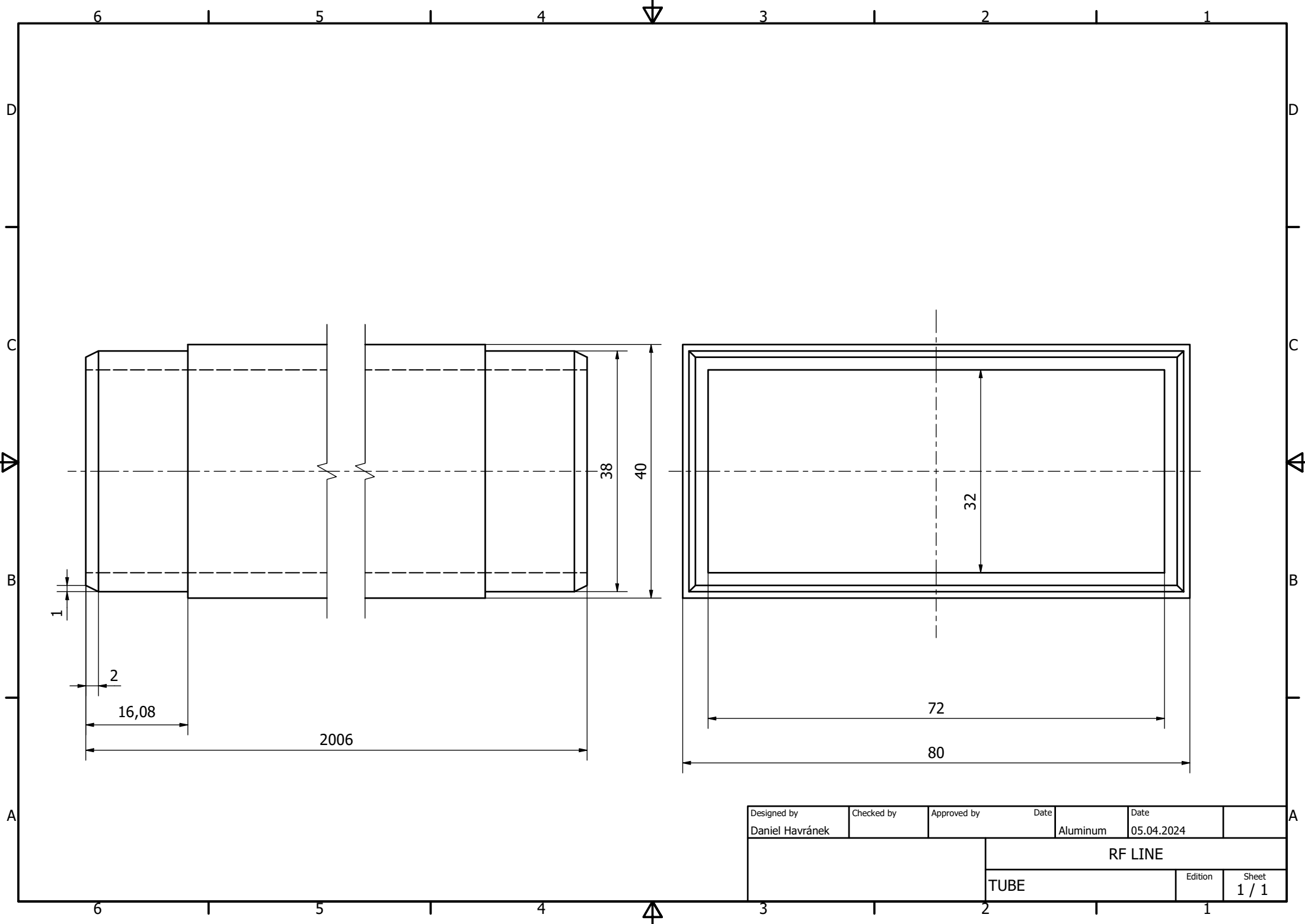
A RF line technical drawings

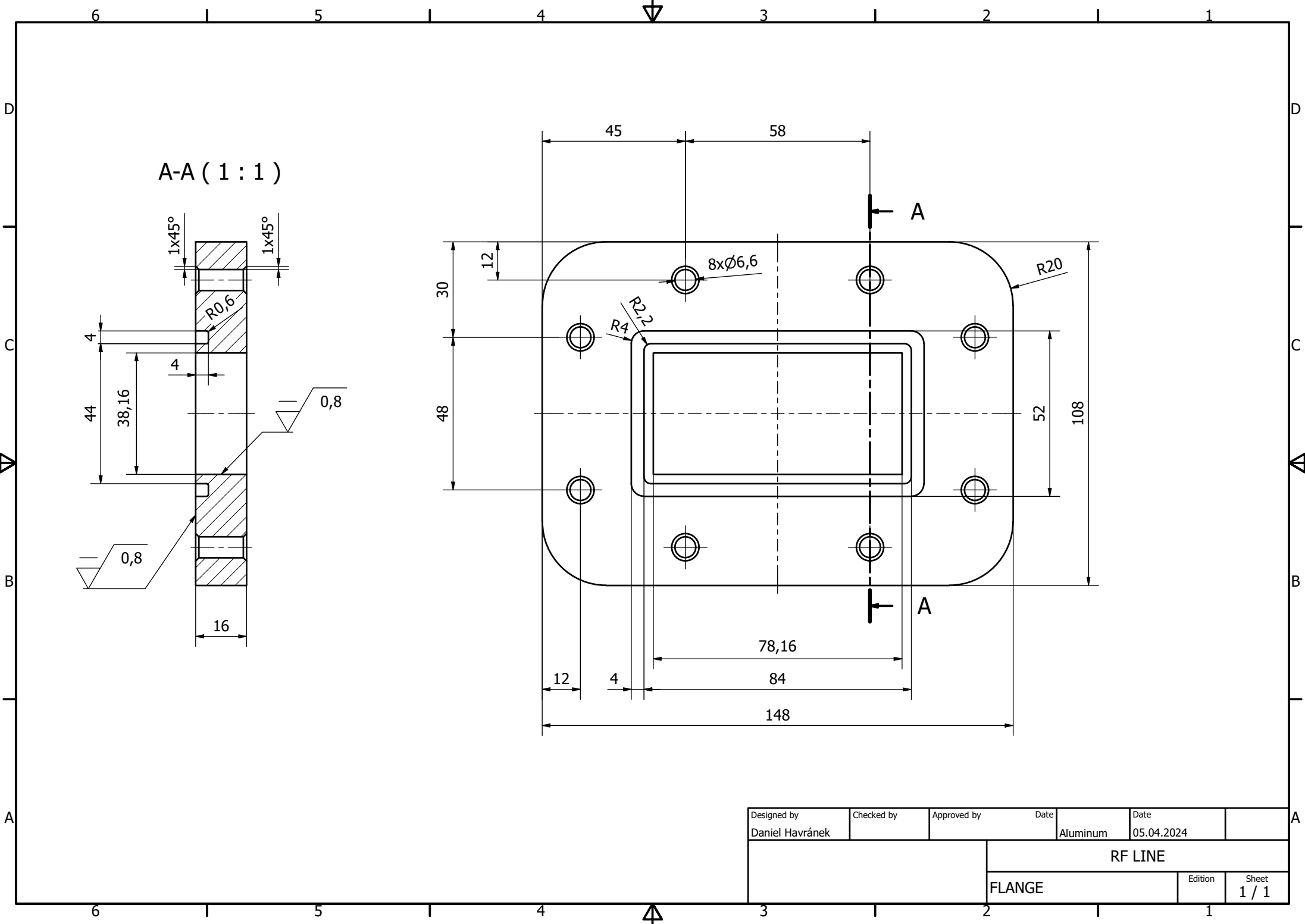


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			STRIP	Edition	Sheet 1 / 1

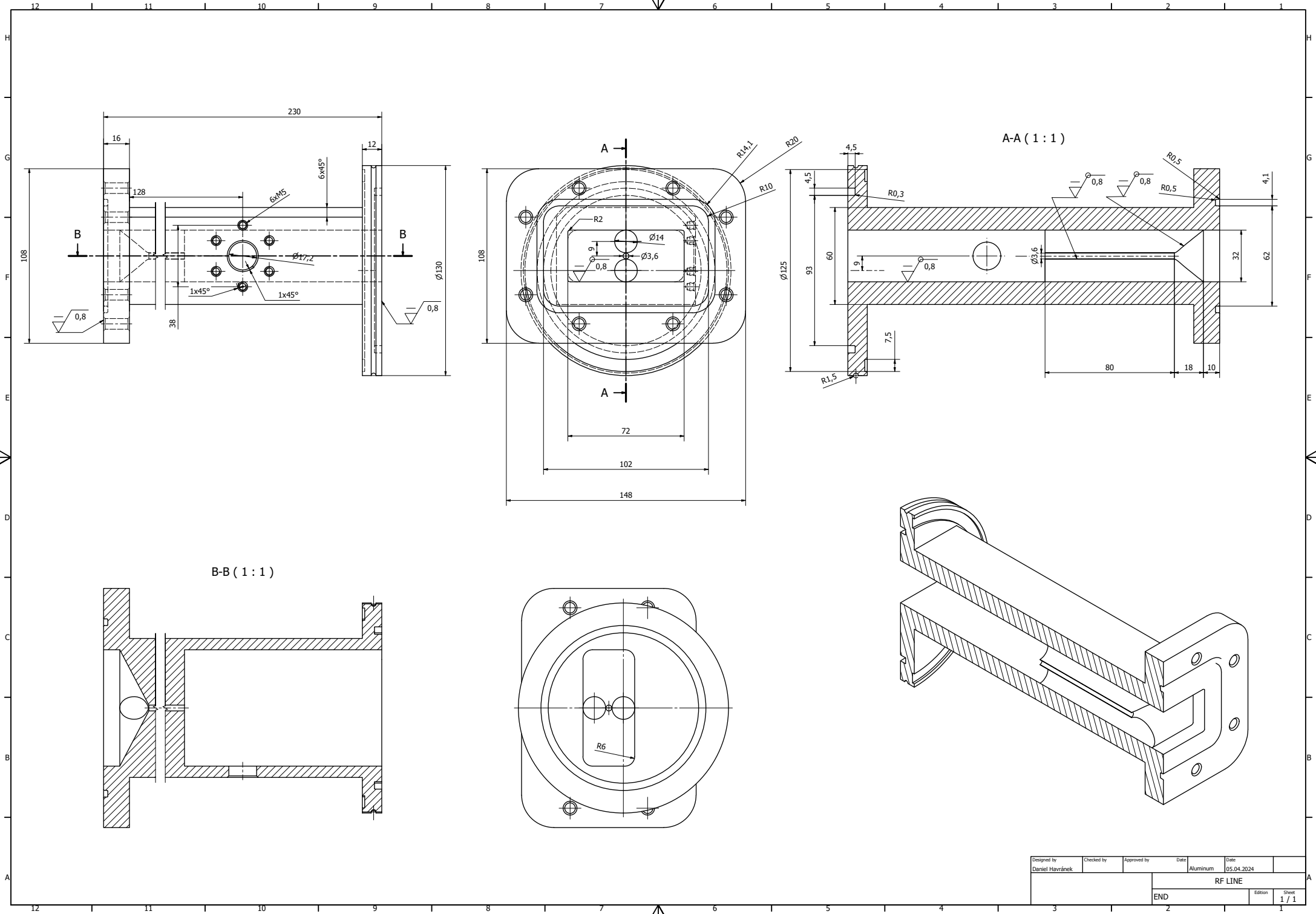
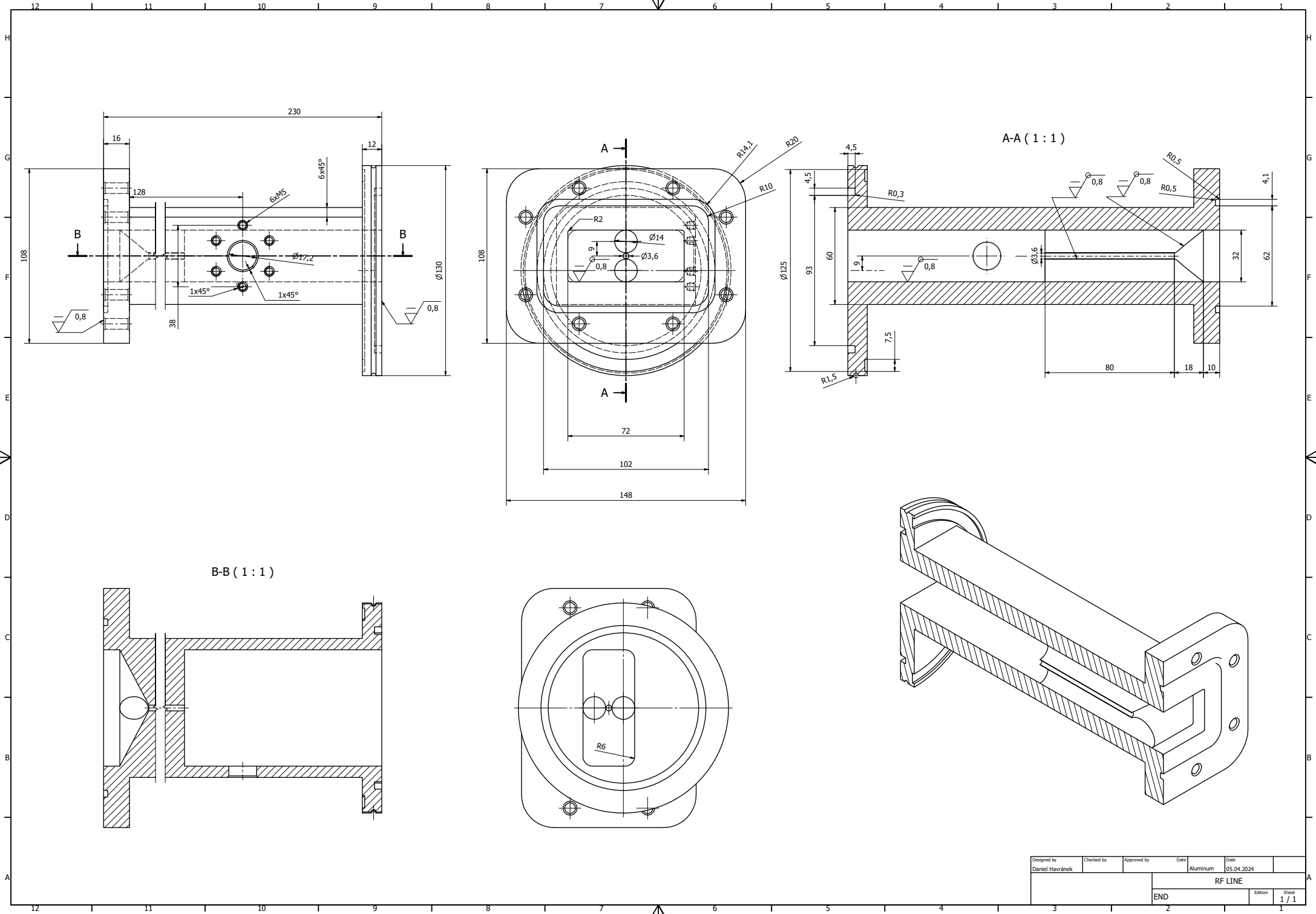
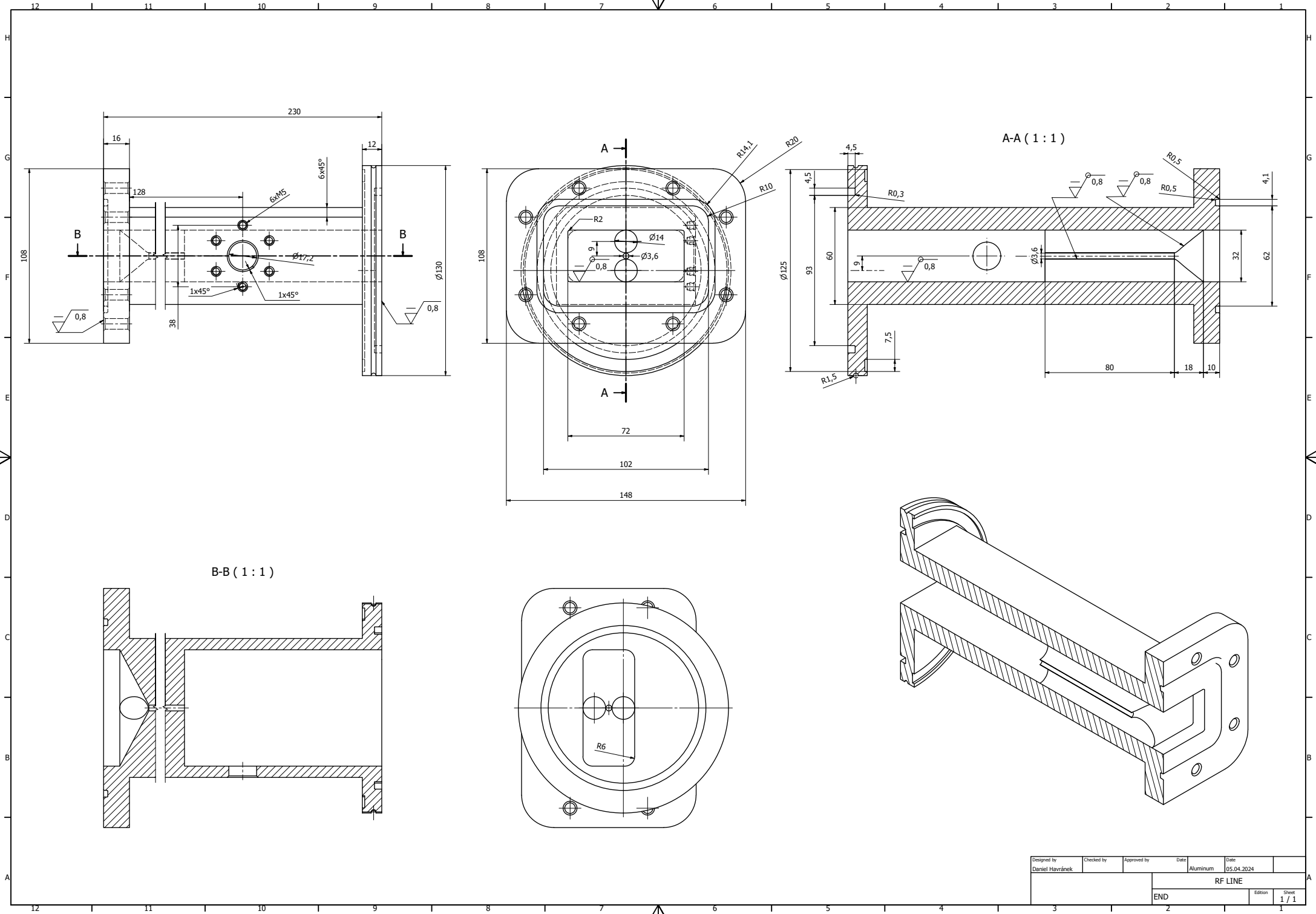
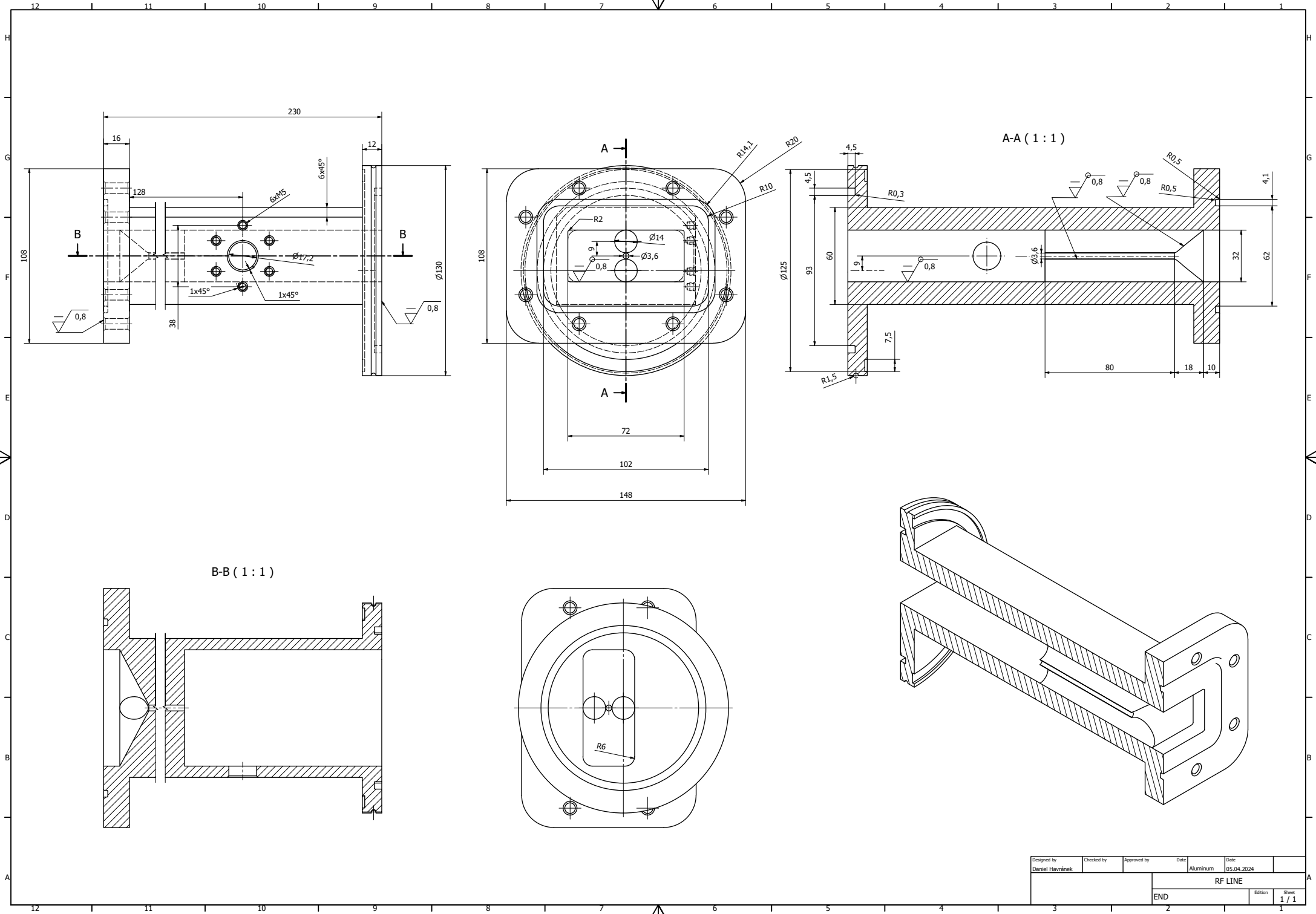
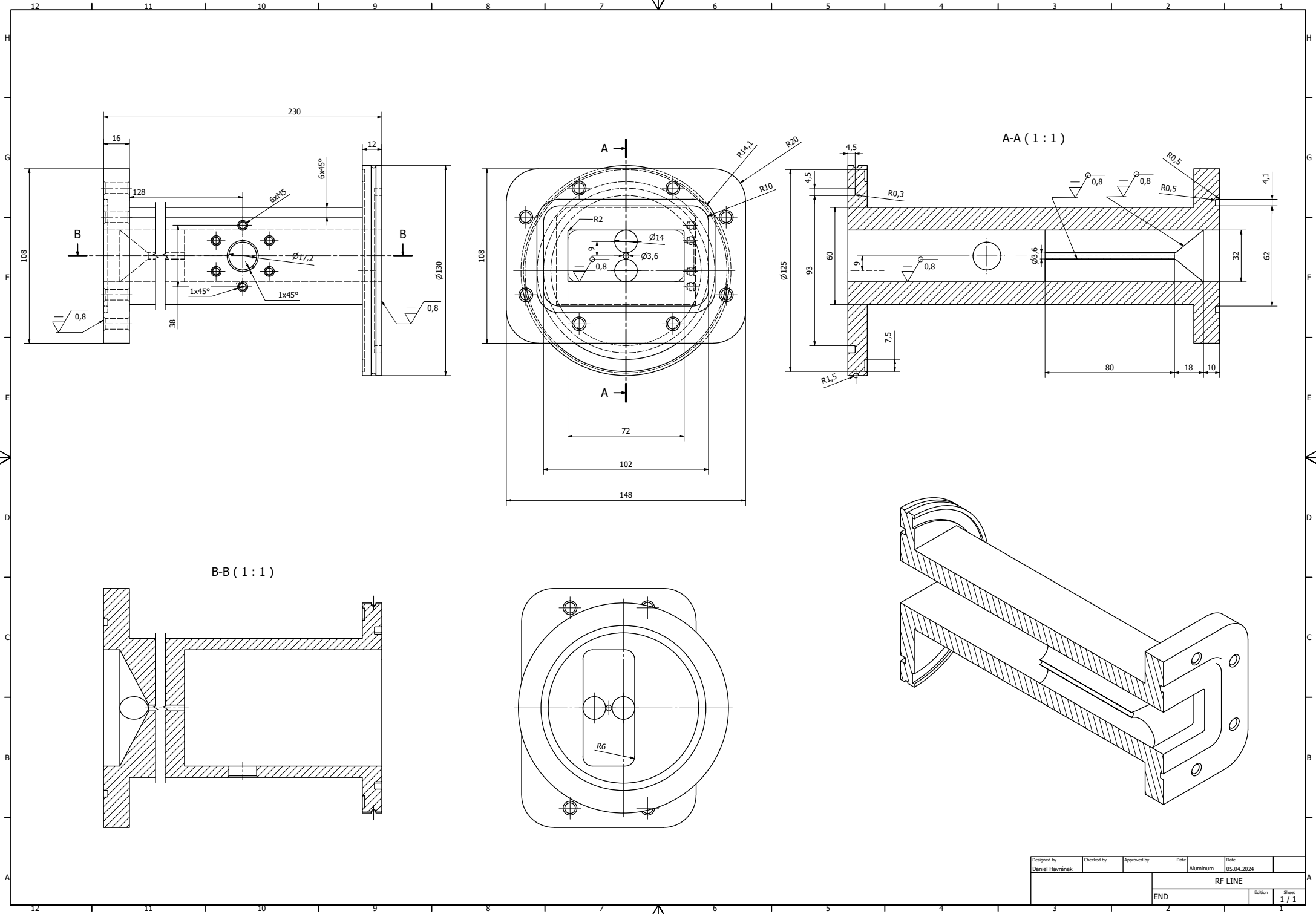
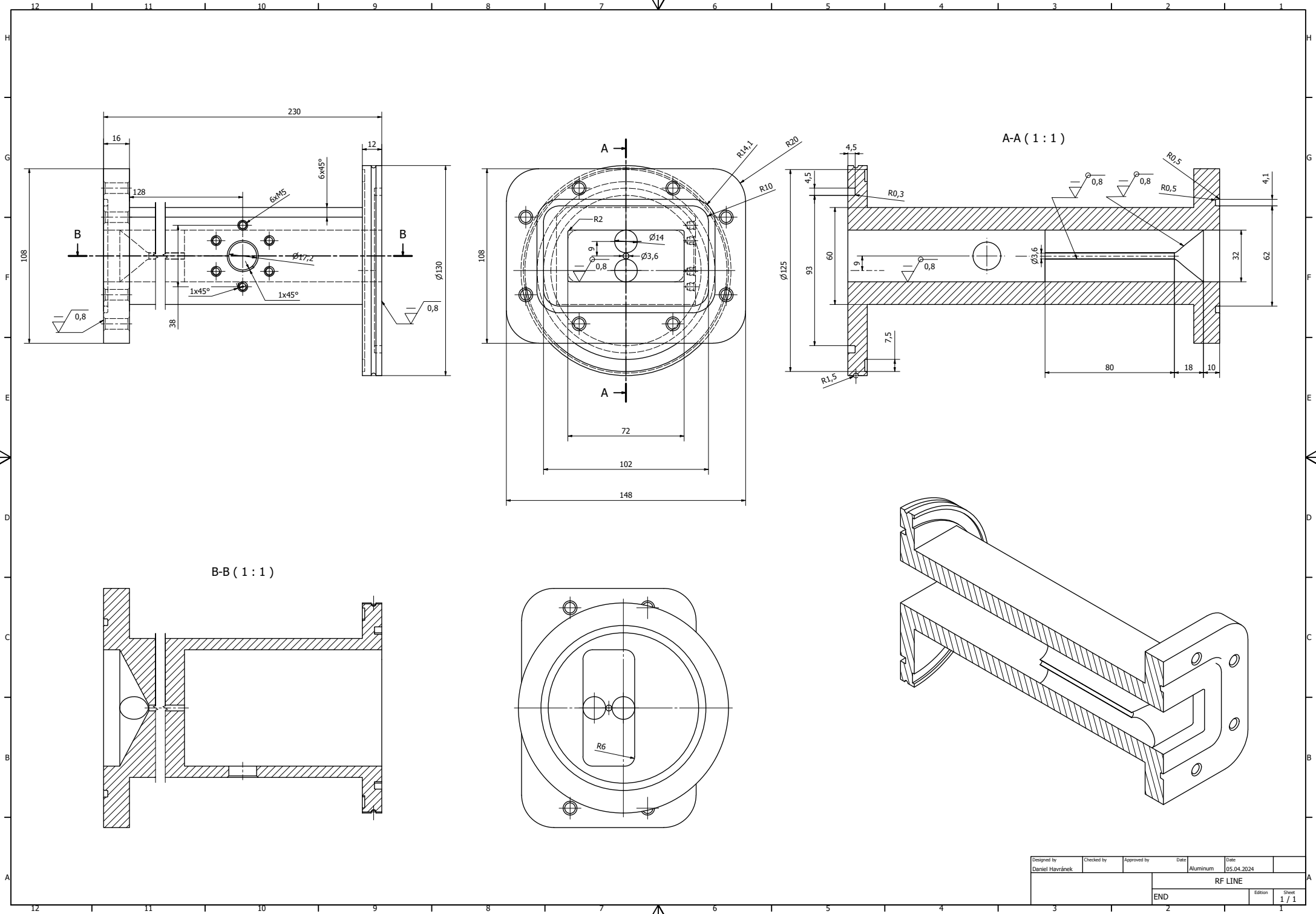


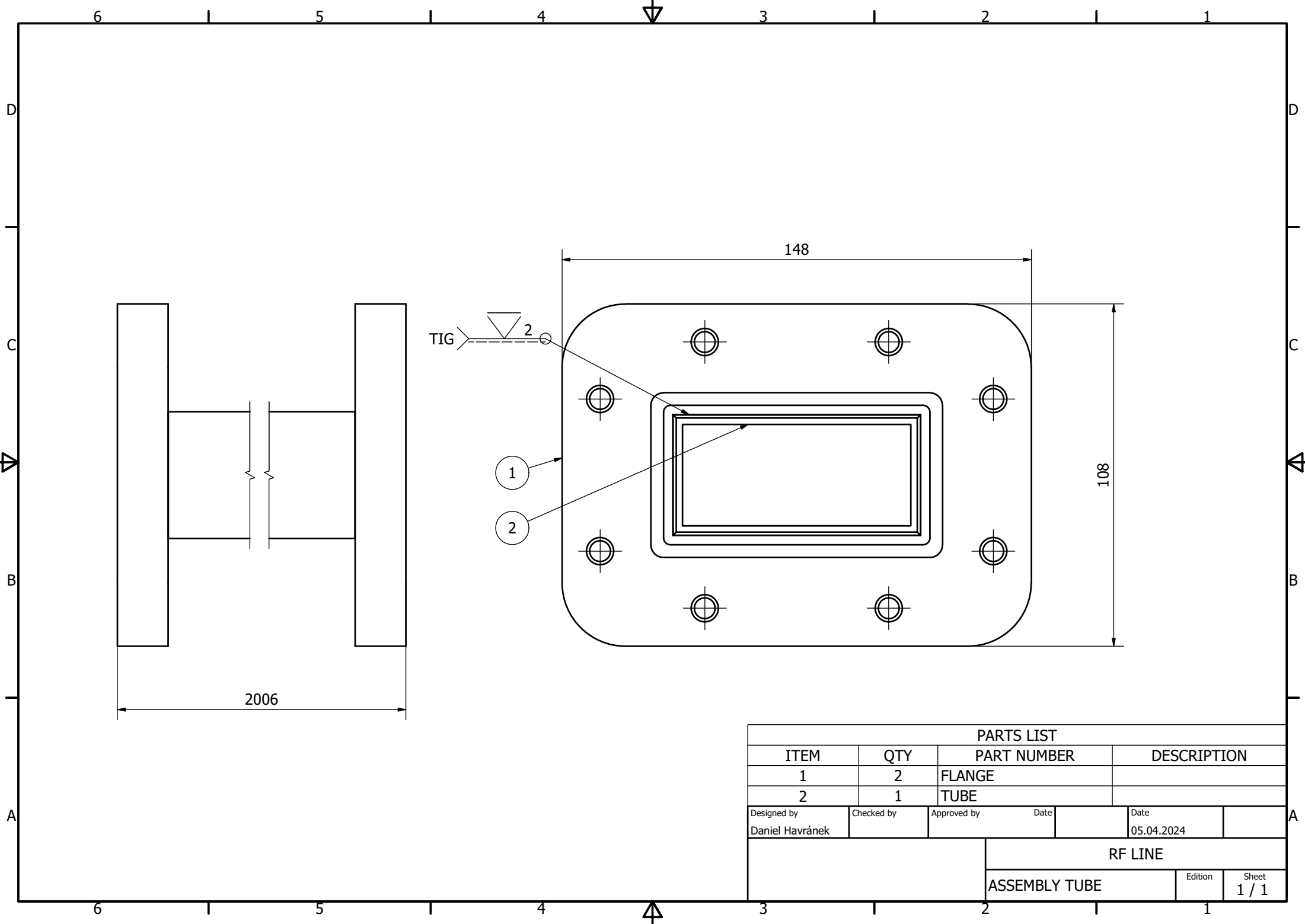
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			SUPPORT	Edition	Sheet 1 / 1



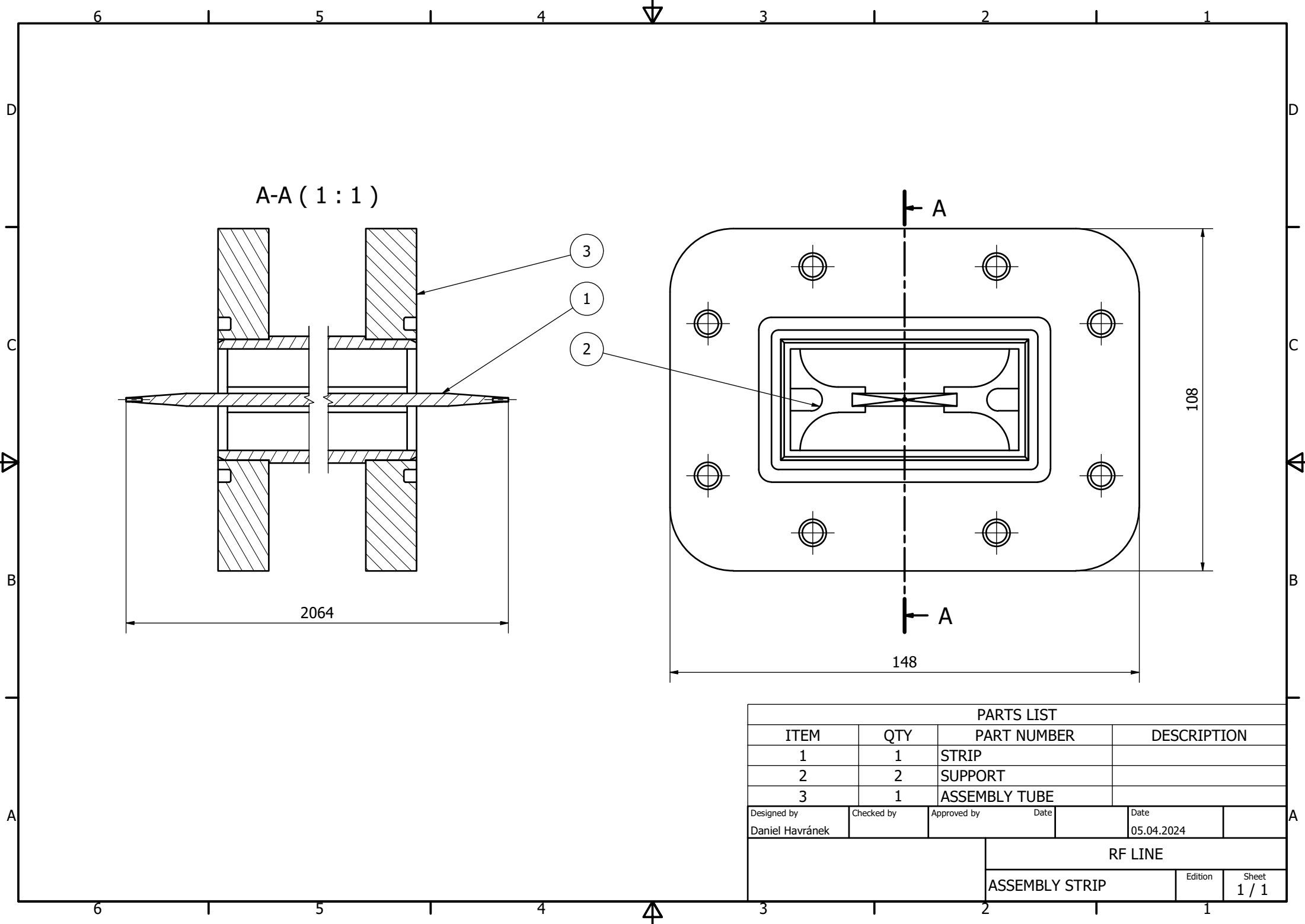


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			RF LINE		
			FLANGE	Edition	Sheet 1 / 1

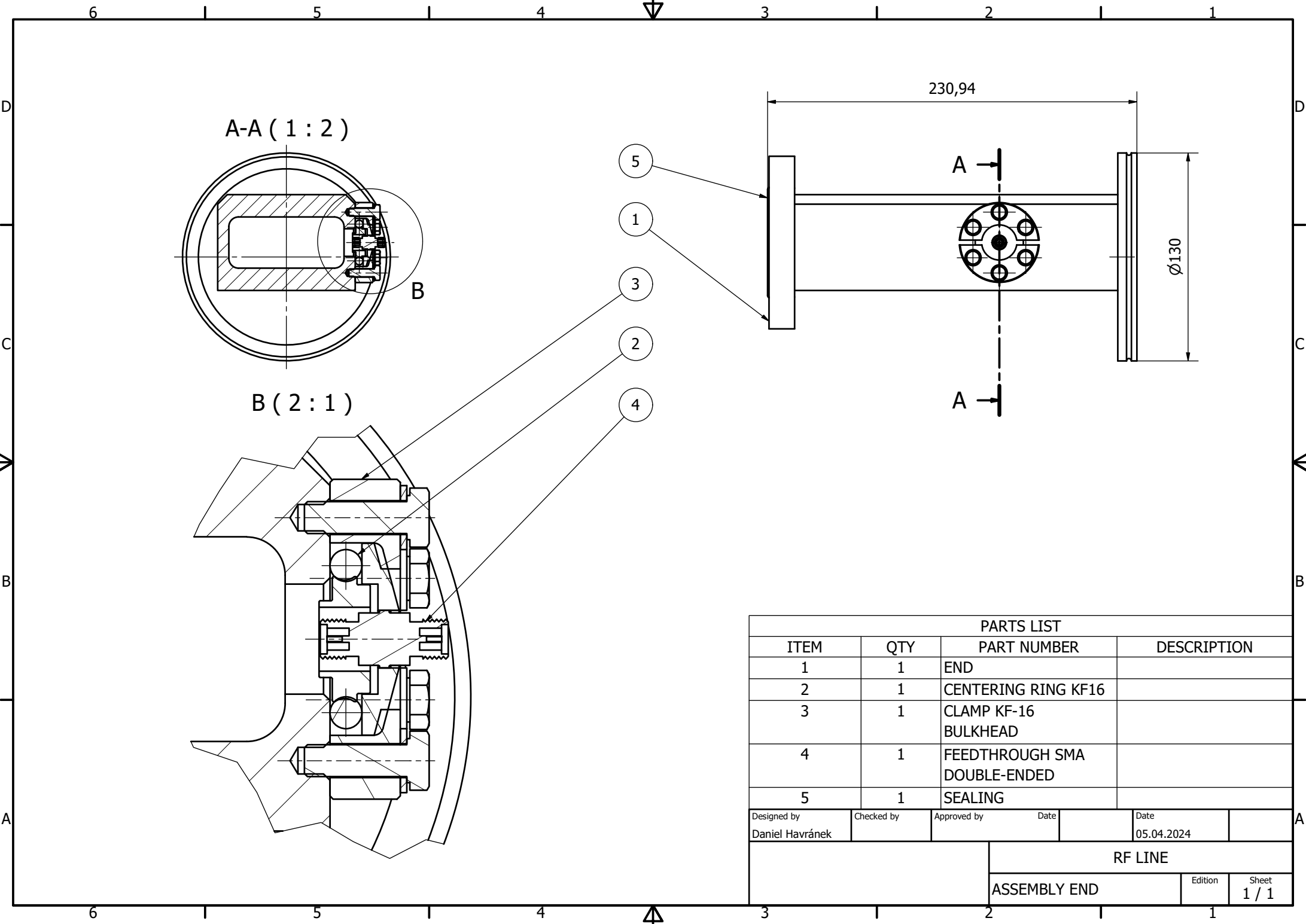
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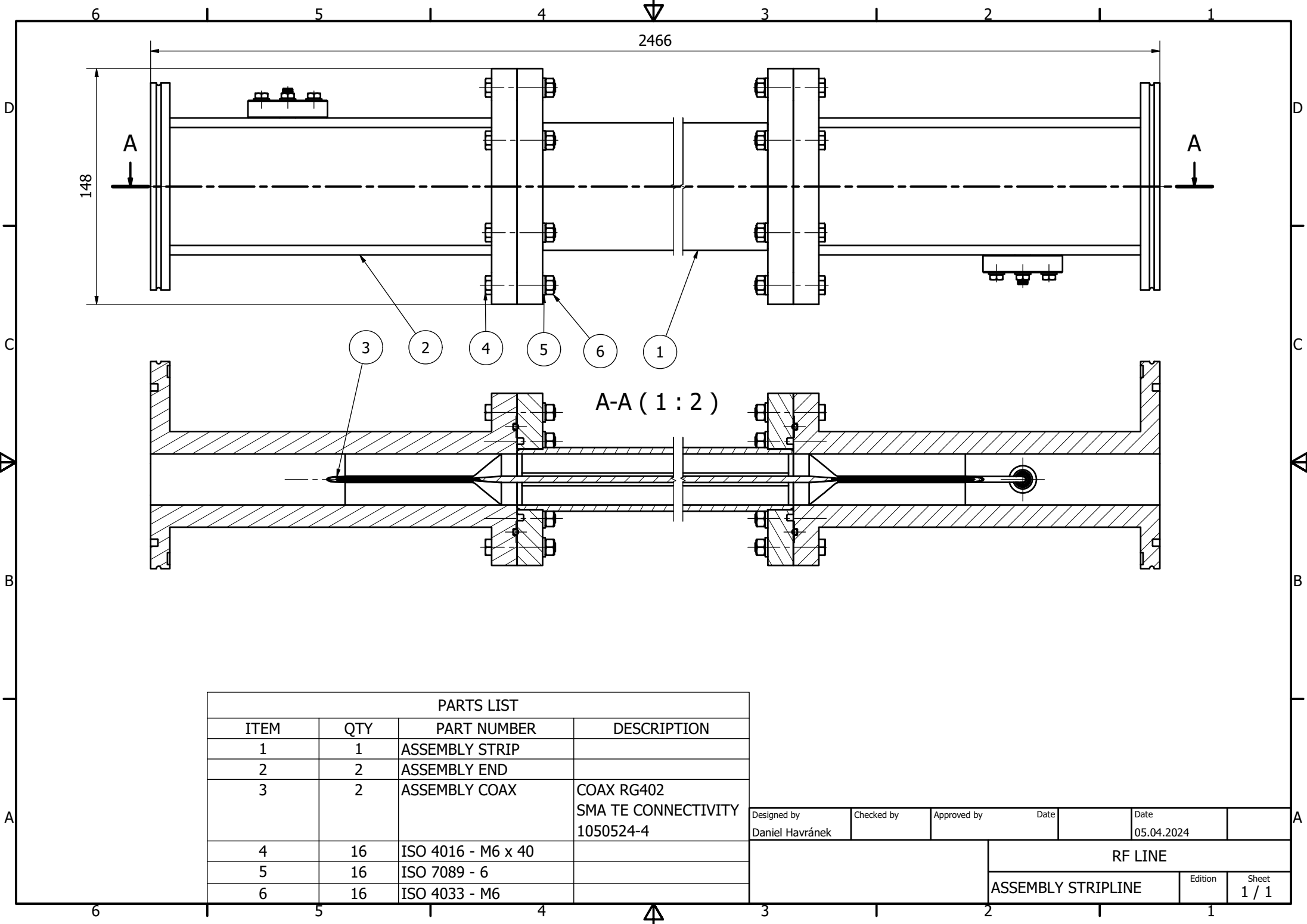


PARTS LIST					
ITEM	QTY	PART NUMBER		DESCRIPTION	
1	2	FLANGE			
2	1	TUBE			
Designed by		Checked by	Approved by	Date	
Daniel Havránek				05.04.2024	
			RF LINE		
			ASSEMBLY TUBE		Sheet 1 / 1



PARTS LIST					
ITEM	QTY	PART NUMBER		DESCRIPTION	
1	1	STRIP			
2	2	SUPPORT			
3	1	ASSEMBLY TUBE			
Designed by Daniel Havránek		Checked by	Approved by	Date 05.04.2024	
			RF LINE		
			ASSEMBLY STRIP		Edition Sheet 1 / 1





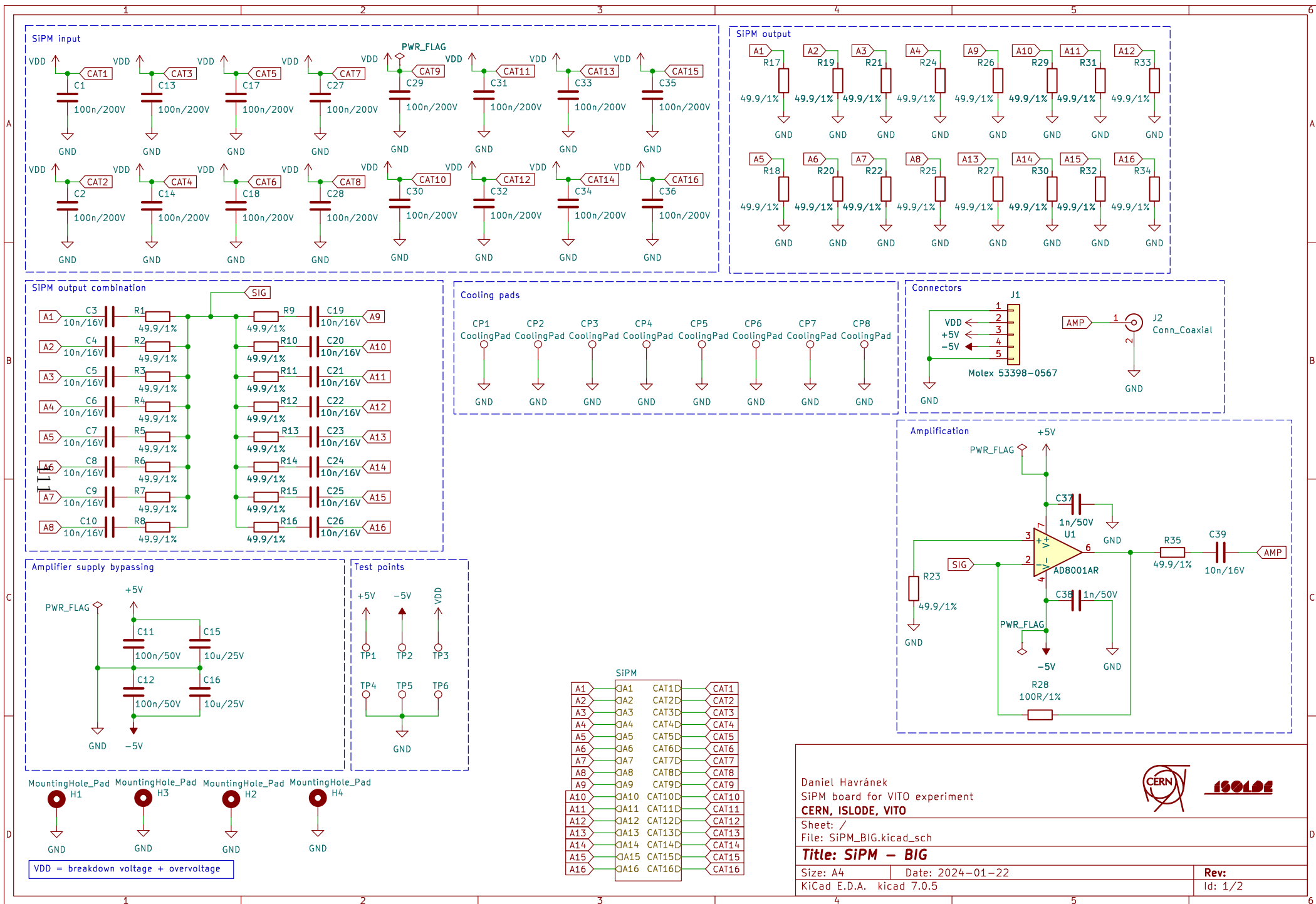
PARTS LIST			
ITEM	QTY	PART NUMBER	DESCRIPTION
1	1	ASSEMBLY STRIP	
2	2	ASSEMBLY END	
3	2	ASSEMBLY COAX	COAX RG402 SMA TE CONNECTIVITY 1050524-4
4	16	ISO 4016 - M6 x 40	
5	16	ISO 7089 - 6	
6	16	ISO 4033 - M6	

Designed by Daniel Havránek	Checked by	Approved by	Date 05.04.2024	
		RF LINE		
		ASSEMBLY STRIPLINE	Edition	Sheet 1 / 1

B Detection electronics for β detector

B.1 Schematic

B.1.1 Detector PCB schematic



Daniel Havránek
SiPM board for VITO experiment
CERN, ISLODE, VITO

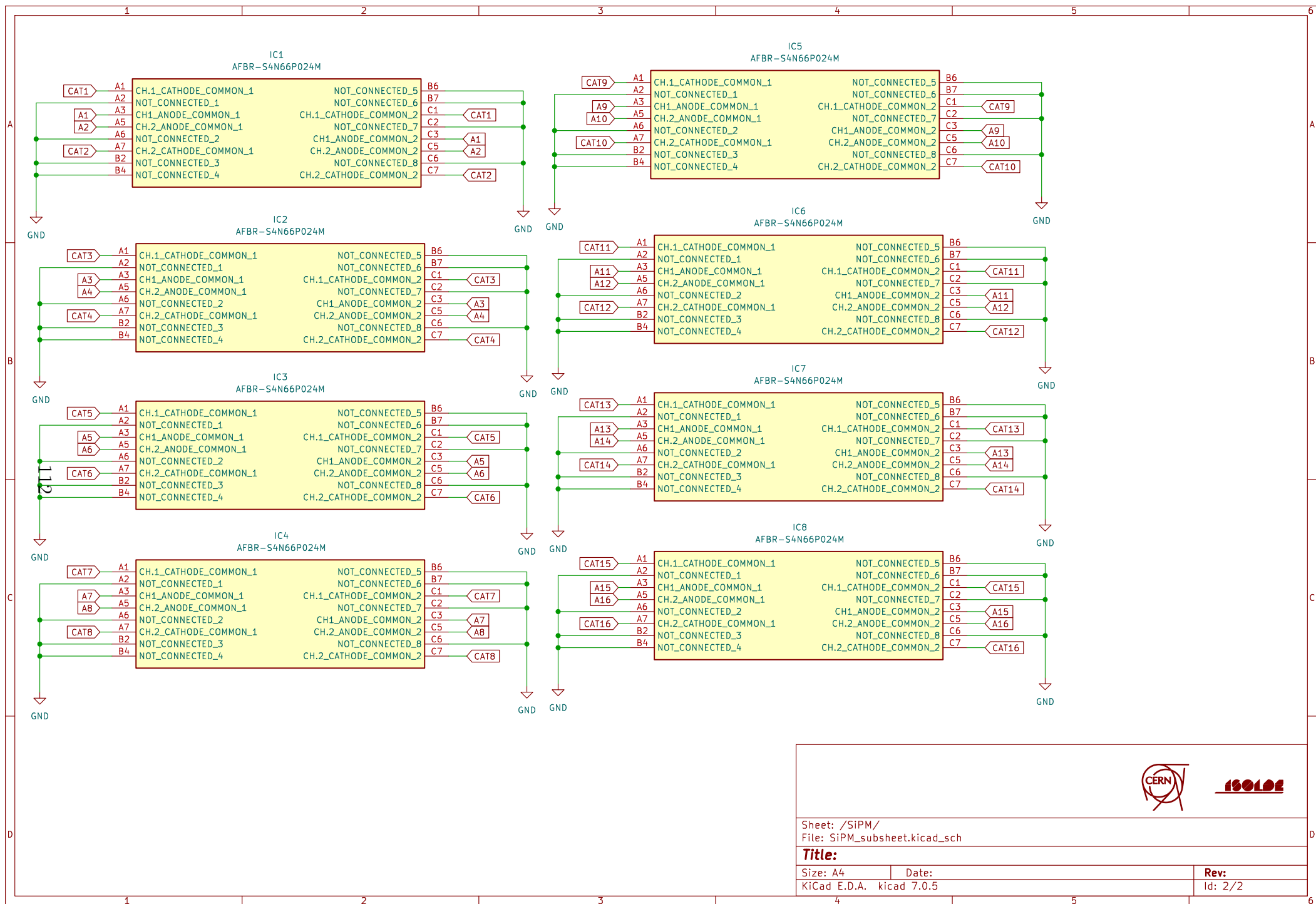


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Title: SiPM – BIG

Size: A4 Date: 2024-01-22
KiCad E.D.A. kicad 7.0.5

Rev:
Id: 1/2



150102

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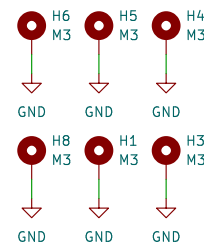
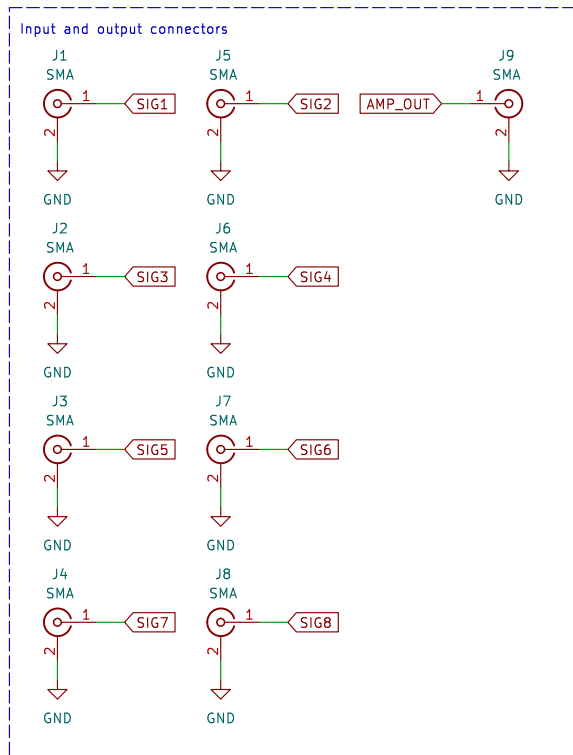
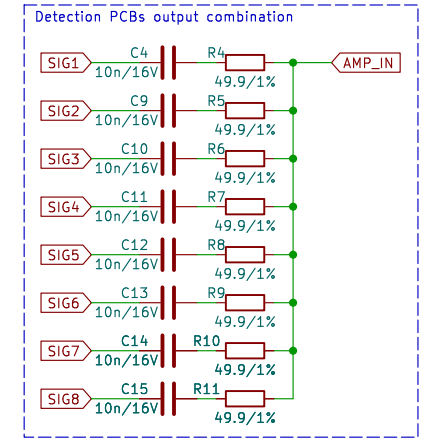
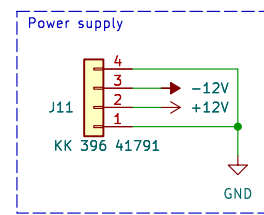
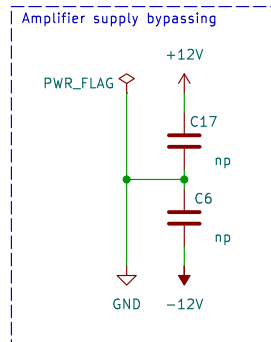
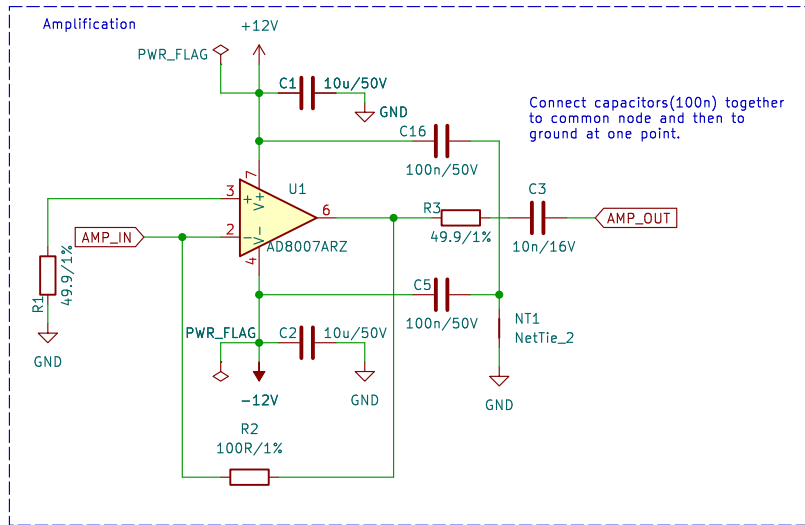
Title:

Size: A4
KiCad E.D.A. kicad 7.0.5

Date:

Rev:
Id: 2/2

B.1.2 Summing PCB schematic

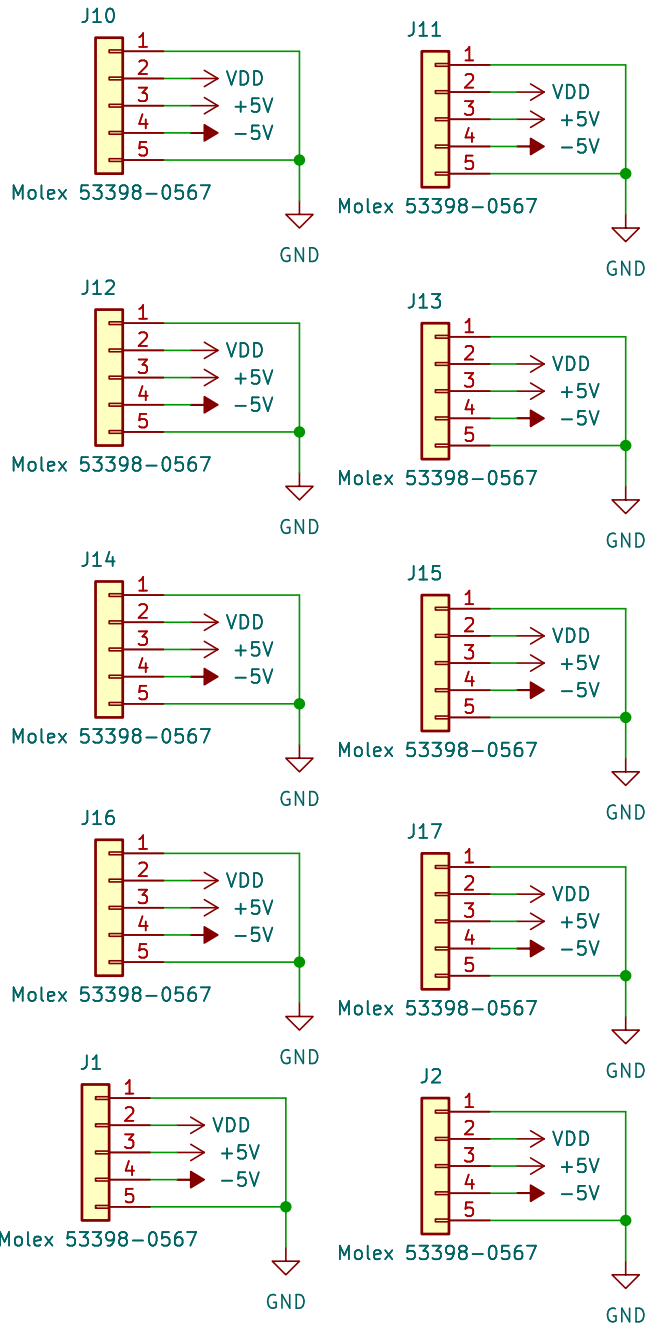




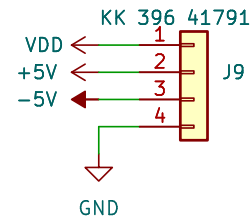

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B.1.3 Power distribution PCB schematic

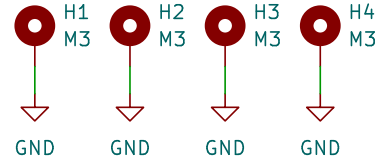
Output connectors



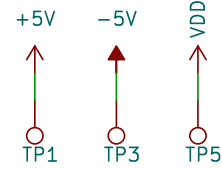
Input connector



Mounting holes



Test points



Sheet: /
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Title: Power distribution board

Size: A5 Date:
KiCad E.D.A. eeschema 7.0.5

Rev:
Id: 1/1

B.2 List of components

Table B.1: List of components - detection PCB

detection PCB - 8x				
Type	Value	Package	Qty	Note
resistor	$49.9\ \Omega$ / 1 %	0402	34	-
resistor	$100\ \Omega$ / 1 %	0402	1	-
capacitor	1 nF / 50 V	0402	2	-
capacitor	10 nF / 16 V	0402	17	-
capacitor	100 nF / 50 V	0402	2	-
capacitor	100 nF / 200 V	1206	16	-
capacitor	10 μ F / 25 V	0805	2	-
amplifier	AD8001	SOIC-8	1	fast, low noise
SiPM	AFBR-S4N66P024M	-	8	-
connector	SAMTEC SMA-J-P-GF-RA-SM1	-	1	coaxial - SMA
connector	Molex 53398-0567	-	1	header

Table B.2: List of components - summing card

summing card - 2x				
Type	Value	Package	Qty	Note
resistor	$49.9\ \Omega$ / 1 %	0603	10	-
resistor	$100\ \Omega$ / 1 %	0603	1	-
capacitor	10 nF / 16 V	0805	9	-
capacitor	100 nF / 50 V	0805	2	-
capacitor	10 μ F / 50 V	0805	2	-
amplifier	AD8007	SOIC-8	1	fast, low noise
connector	Molex 73391-0080	-	9	coaxial - SMA
connector	Molex KK 396 41791	-	1	header

Table B.3: List of components - power distribution PCB

power distribution PCB - 2x				
Type	Value	Package	Qty	Note
connector	Molex 53398-0567	-	34	header
connector	Molex KK 396 41791	-	1	header

Table B.4: List of components - filter

Filter - 2x				
Type	Value	Package	Qty	Note
resistor	50 Ω / 0.1 %	1206	1	-
capacitor	47 pF / 50 V	1206	1	-
connector	Molex 73391-0080	-	2	coaxial - SMA

B.3 β detector electronics graphs

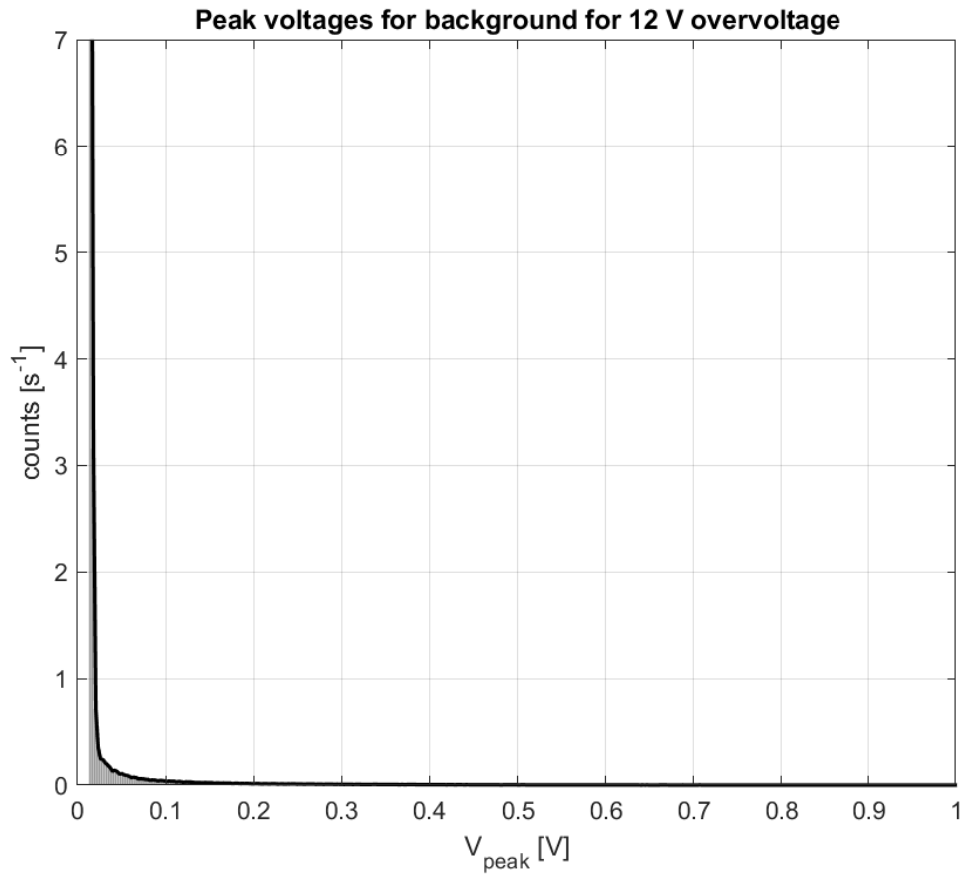


Fig. B.1: Measured peak voltage for background (no source) for different overvoltage.

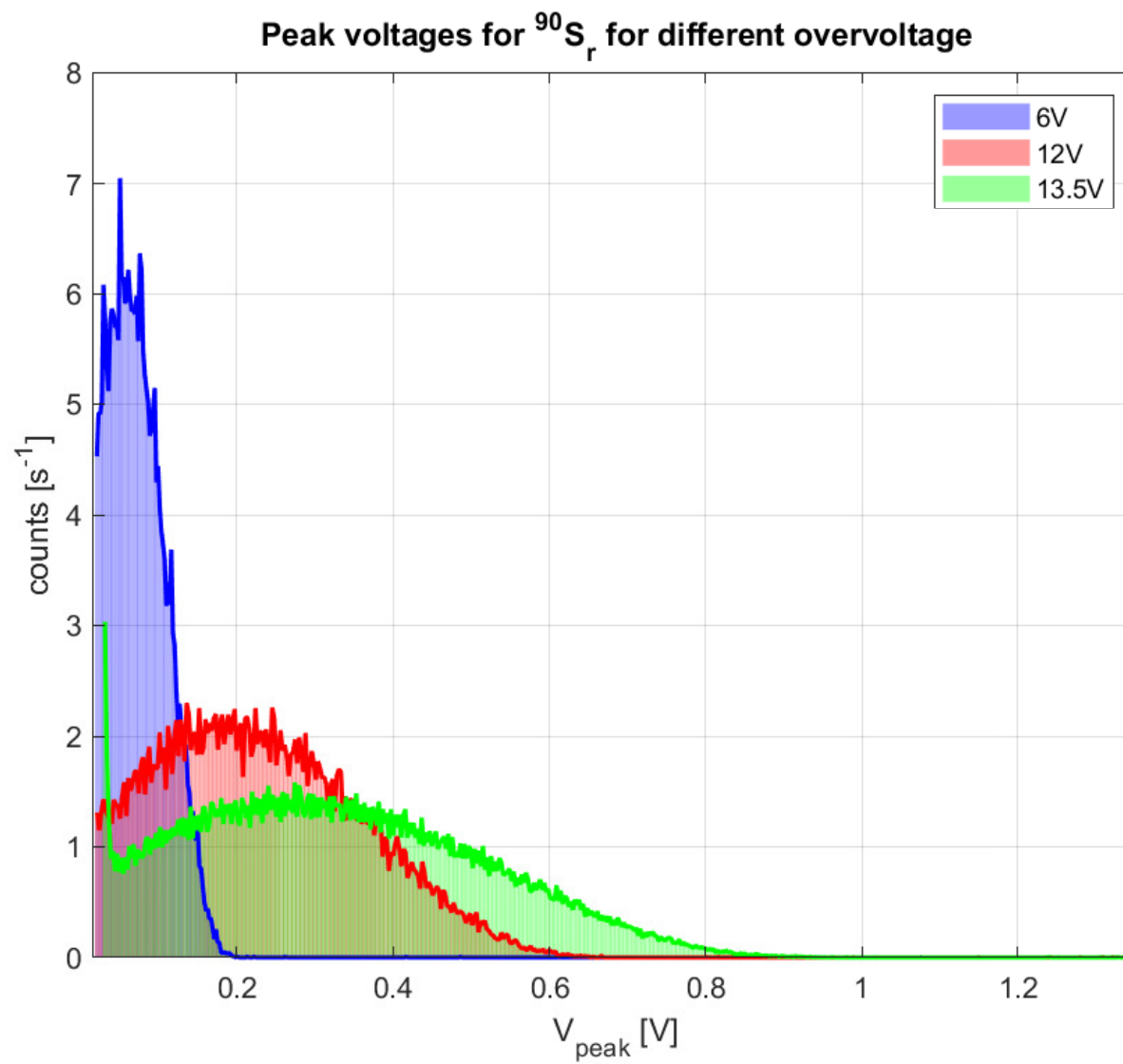


Fig. B.2: Measured peak voltage for ^{90}Sr for different overvoltage.

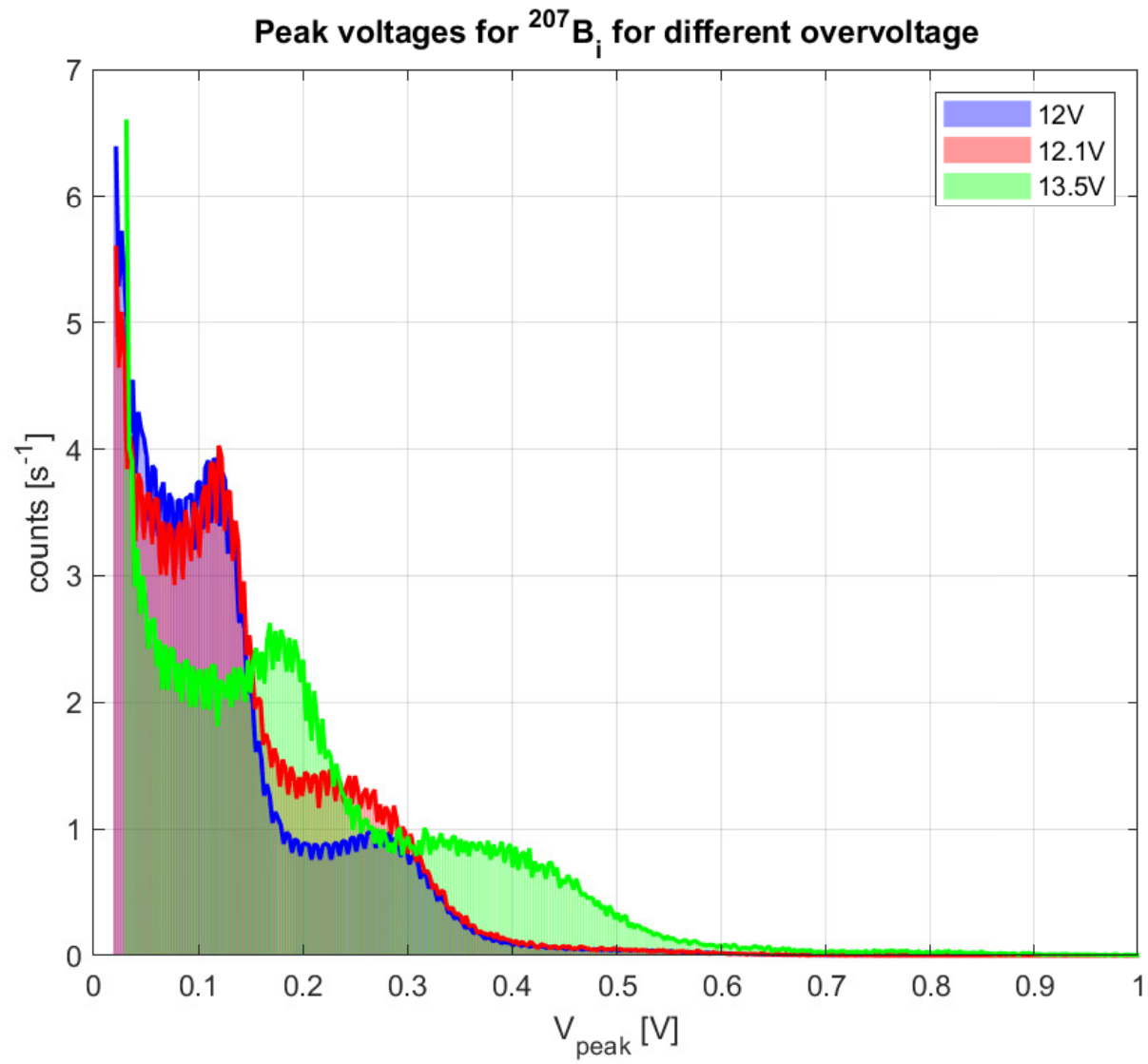


Fig. B.3: Measured peak voltage for ^{207}Bi for different overvoltage.

C Images

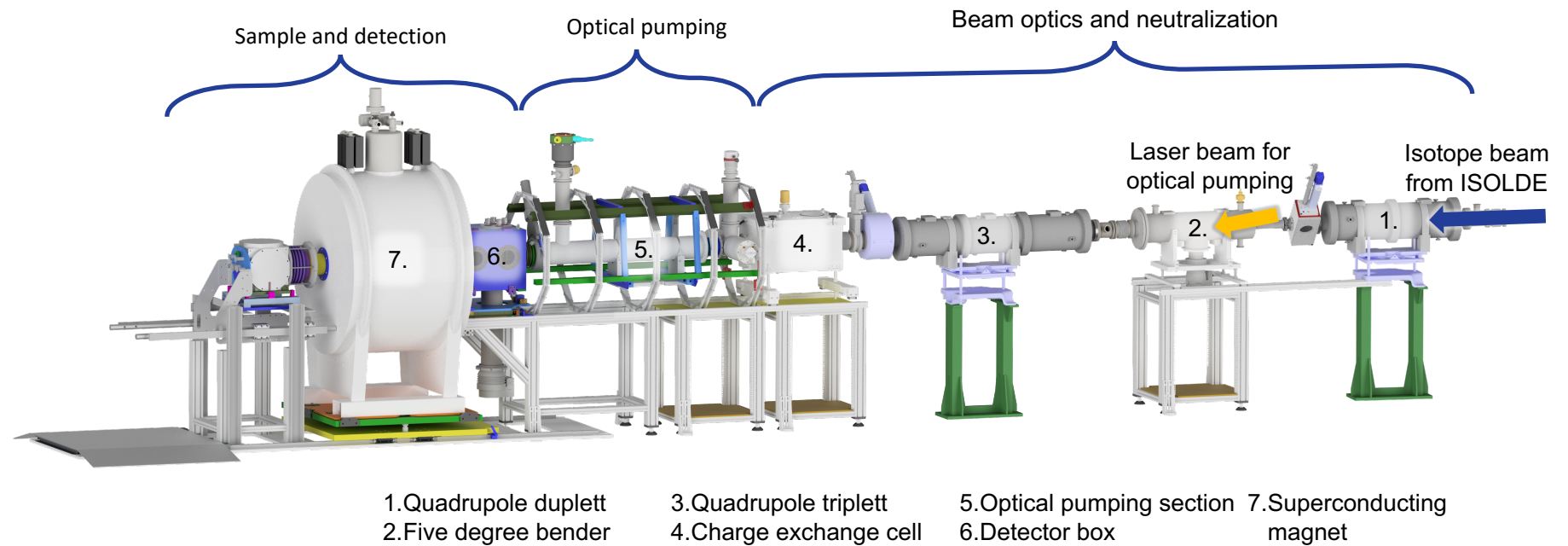


Fig. C.1: VITO experiment setup.

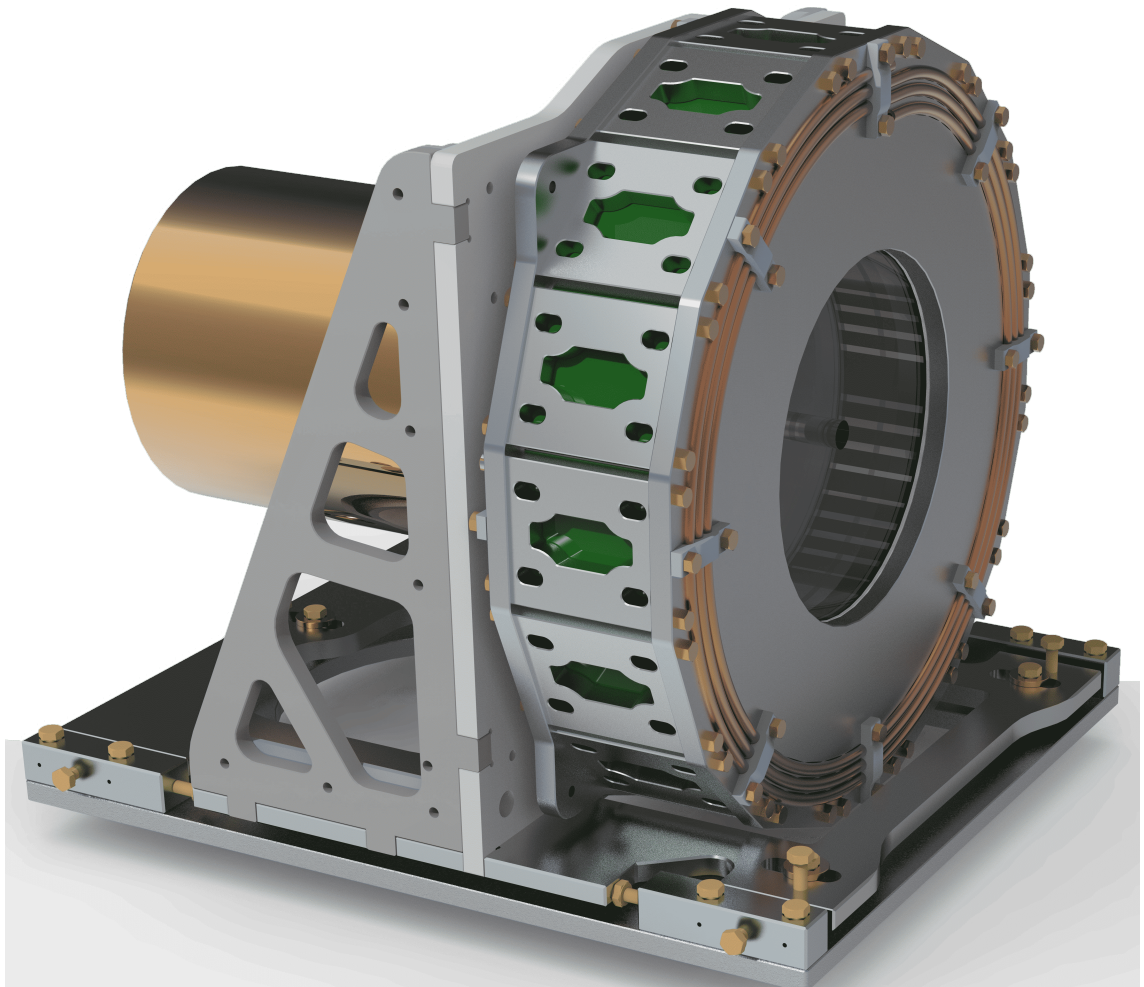


Fig. C.2: Big detector - preliminary.

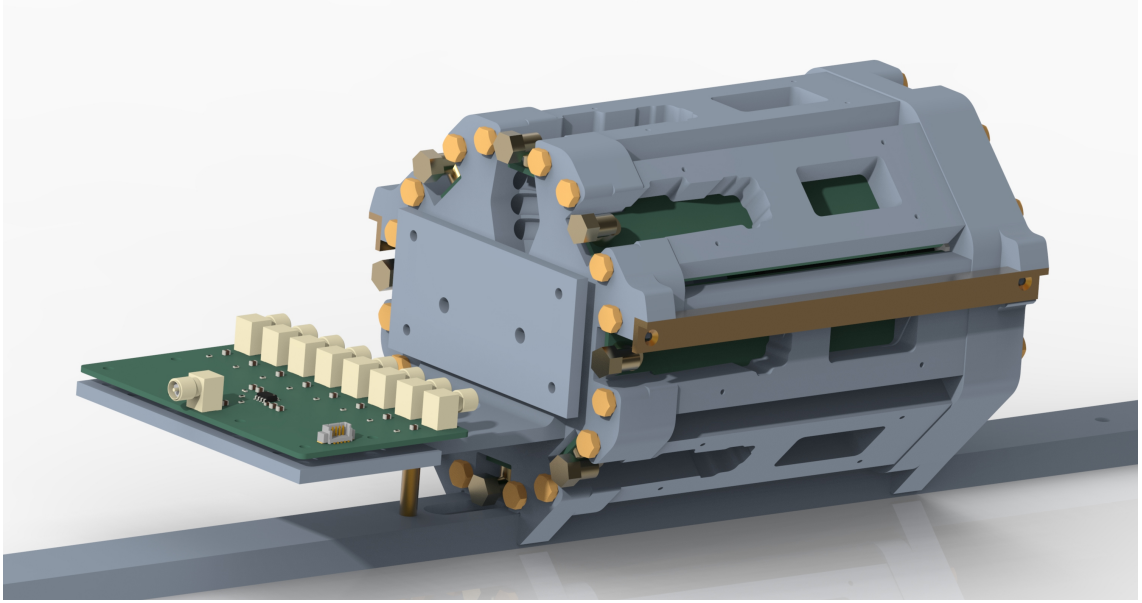


Fig. C.3: Small detector (front) - preliminary.

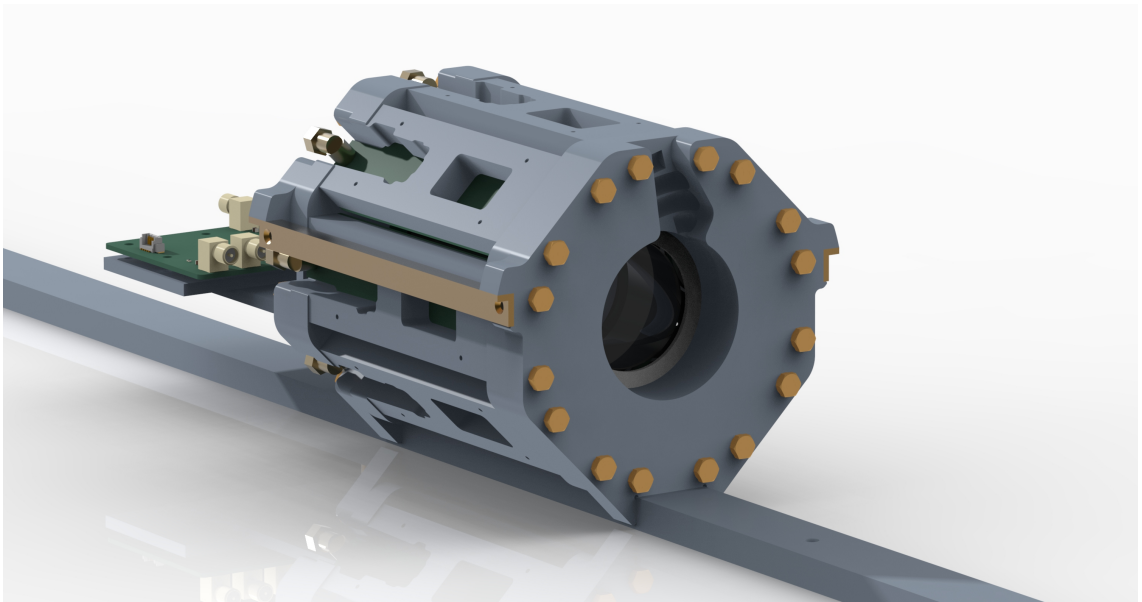


Fig. C.4: Small detector (back) - preliminary.