

Master Thesis

Setup and characterization of a silicon photonic transmitter system

Aufbau und Charakterisierung eines
siliziumphotonischen Transmittersystems

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Karlsruhe der 11.02.2026

Heinrich Mayer

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List of abbreviations

Abbreviation	Meaning
HL	High-Luminosity
LHC	Large Hadron Collider
EO	Electro-Optic
WDM	Wavelength-Division Multiplexing
cw	Continuous Wave
MUX	MUltipleXer
DEMUX	DEMultipleXer
SMF	Single-Mode Fiber
DAQ	Data AQuisition
SOI	Silicon-On-Insulator
PDE	Plasma Dispersion Effect
TID	Total Ionizing Dose
TIR	Total Internal Reflection
TE	Transverse Electric
MMF	Multi-Mode Fiber
MFD	Mode Field Diameter
PM	Polarization-Maintaining
PIC	Photonic Integrated Circuit
GC	Grating Coupler
MZM	Mach-Zehnder Modulator
RM	Ring Modulator
BOX	Buried OXide
RVB	ReVerse Bias
FWB	ForWard Bias
HF	High Frequency
SiPh	Silicon Photonic(s)
FSR	Free Spectral Range
HEP	High Energy Physics
PDK	Process Design Kit
EOE	Electro-Optical Electric
SE	Shallow Etch
DE	Deep Etch
EOE	Electrical-Optical-Electrical
SMU	Source/Measure Unit
OPM	Optical Power Meter
FCW	Focusing Wideband (referring to a grating coupler)
CPW	CoPlanar Waveguide
GCPW	Grounden CoPlanar Waveguide
MMI	Multi-Mode Interferometer
DMM	Digital MultiMeter

RF	Radio Frequency
RIA	Radiation-Induced Attenuation
SRF	Self-Resonant Frequency

1 Introduction

Electronics within particle detectors of particle collision experiments are often exposed to harsh X-ray and particle radiation. This is especially true close to the particle beam pipe at the center of detectors such as the ATLAS or CMS experiments at the LHC at CERN [1]. With the upcoming High-Luminosity (HL) Upgrade to the LHC [2] the radiation will increase significantly. In combination with higher resolution demands and the goals to move more data processing off-detector as well as to remove mass, the readout systems need to be upgraded with newer technologies to facilitate the resulting higher data rates.

One such proposed technology is based on electro-optic (EO) modulators. Opposed to the directly modulated laser diodes currently utilized, optical data transmission systems based on EO modulators use an external light source whose light is then modulated to achieve signal transmission. In combination with wavelength-division multiplexing (WDM), which enables the use of one single optical fiber for several data channels by combining several continuous wave (cw) monochromatic laser outputs acting as individual data channels, multi-modulator setups can achieve aggregated data rates in the terabit-per-second range [3]. A schematic of such a WDM based data transmission system is shown in figure 1.1.

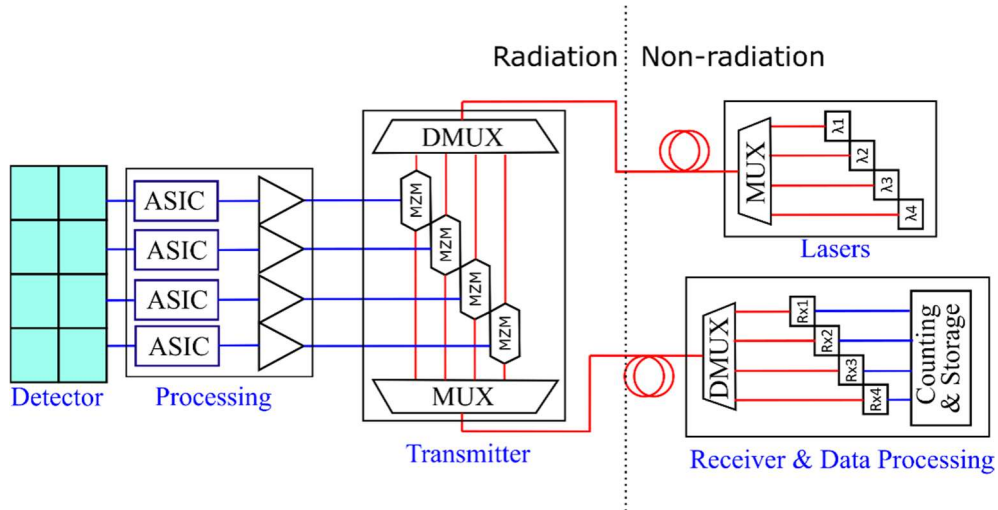


Figure 1.1: Schematic of the proposed generic high-speed transmitter design with the irradiated and the non-irradiated part (taken from [4]).

Several cw monochromatic laser outputs are fed into a multiplexer (MUX). Therefore, all wavelengths can be sent through a single optical single-mode fiber (SMF) to the detector. Detector-side the wavelengths are separated by a demultiplexer (DEMUX) and each is led through an EO-modulator, which encodes the electrical signal from the sensor and read-out chips onto the cw input. The individual wavelengths are then led into a MUX and routed through another SMF to the counting room. There a DEMUX separates the wavelengths a final time and the individual encoded signals are captured by a data acquisition (DAQ) backend to undergo further processing.

Modern EO-modulators, based on silicon-on-insulator (SOI) technology, fabricated in readily available CMOS-like processes, are the most promising avenue and are pursued for online detector applications. Using the plasma dispersion effect (PDE) they offer very high modulation and therefore signal encoding

speeds. Additionally, with this technology driver circuits, waveguides and optical (DE)MUXs can be implemented next to the modulators on one single chip, therefore offering a monolithic, space-efficient approach. Also, radiation hard designs that can withstand the conditions of the detector environment can be realized relatively easily.

The radiation hardness of such integrated designs with focus on the modulators is currently investigated by several groups [5,6]. Modulators lose their modulation capabilities after a certain total ionizing dose (TID). To compensate for this, either the TID ceiling until loss of modulation capability must be increased by changing aspects of the design or a so-called annealing step must take place, so that the functionality is restored.

The primary objective of this thesis is to set up and characterize single modulator data transmission systems using the COTTONTAIL chip [7] under in-situ conditions to investigate and quantify their behavior under X-ray irradiation. The COTTONTAIL schip is the most recent SOI WDM-based chip of the group [8], containing isolated test structures and fully integrated multichannel data transmission systems. Several groups are currently working on the advancement of technology regarding its radiation hardness. The approaches differ by the wavelength range used and the distance between the single modulation wavelengths of that range [9-11]. To enable the irradiation measurements, several preparations and optimizations had to be performed to the packaging and interconnection technology previously used by the group [12]. Specifically, fan-out PCBs for the high-frequency broadband electrical connection of the single modulators via wire bonds were developed to connect the devices to a signal generator. Furthermore, aspects of the fiber-to-chip coupling for optical signal transmissions were improved to enable the measurements. Finally, software for the main and supporting measurements was created.

The remainder of this thesis is structured as follows: Chapter 2 presents the necessary theoretical background of opto-electronics, radiation damage and associated electronics. Chapter 3 introduces the specific structures and components on the COTTONTAIL chip investigated in this work. Chapter 4 details the packaging and interconnection technology developed, covering fiber-to-chip coupling (Section 4.2), the design of the fan-out PCBs (Section 4.4) and the final assembly of in-situ measurement samples (Section 4.5). This preparation leads to Chapter 5, which presents X-ray irradiation measurements and characterization results. Finally, this thesis concludes with a summary and outlook in Chapter 6.

2 Theoretical background

2.1 Optics and opto-electronics

Mathematically, the propagation of a single wave along the z-axis can be written as [13]

$$E = E_0 e^{i(k_z z \pm \omega t)} \text{ and} \quad (2.1)$$

$$H = H_0 e^{i(kz \pm \omega t)}. \quad (2.2)$$

$|k| = 2\pi/\lambda$ is the propagation constant in the propagation direction and quantifies the spatial phase change. ω is the angular velocity, E is the electric field with its magnitude of E_0 , whereas H and H_0 are the corresponding magnetic field quantities. To relate electromagnetic fields to currents and charge distributions, Maxwell's equations are needed.

Ohm's law $j = \sigma E$ with the conductivity σ enables us to derive the wave equations for non-conductive materials without access charges out of the Maxwell equations for [13]. They are

$$\nabla^2 E = \mu_m \epsilon_m \frac{\partial^2 E}{\partial t^2} \quad (2.3)$$

for the electric field and

$$\nabla^2 H = \mu_m \epsilon_m \frac{\partial^2 H}{\partial t^2} \quad (2.4)$$

for the magnetic field. ϵ_m is the permittivity of a medium representing its ability to be polarized. Similarly, the permeability μ_m represents the ability of a medium to be magnetized. They describe the propagation of electromagnetic waves. Comparing both wave equations to the general form of a wave equation yields the phase velocity [14]

$$v = \frac{1}{\sqrt{\mu_m \epsilon_m}} = \frac{\omega}{k}. \quad (2.5)$$

The refractive index n relates v , which is the propagation speed of the wavefront, to the speed of light $c = 1/\sqrt{(\mu_0 \epsilon_0)}$ by

$$n = \frac{c}{v} = \pm \sqrt{\frac{\mu_m \epsilon_m}{\mu_0 \epsilon_0}} = \pm \sqrt{\mu_r \epsilon_r} = \frac{k}{k_0}. \quad (2.6)$$

μ_0 and ϵ_0 are the values of μ_m and ϵ_m in vacuum; therefore, μ_r and ϵ_r are their respective relative values.

Now we consider an incident light beam E_i coming from a medium with a refractive index of n_1 hitting the interface to another medium with n_2 . This is depicted in figure 2.1. The beam hits the interface at an angle of θ_1 . There it is reflected at θ_1 as E_r or refracted into the other media at θ_2 as E_t . This change in angle occurs due to the change in phase velocity and ensures the continuity of wave fronts, which is demanded by the law of energy conservation.

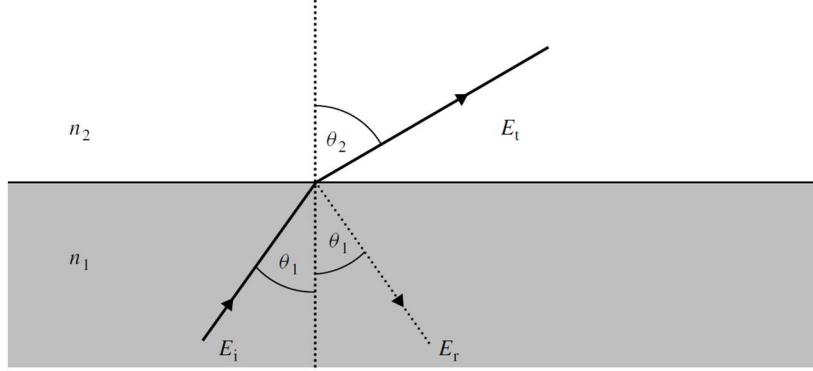


Figure 2.1: Refraction of a light ray at an interface between two media. The incident ray E_i is split into a transmitted (E_t) and a reflected (E_r) ray (taken from [12] as an adaption from [15]).

The relation between θ_1 and θ_2 is given by Snell's law

$$n_1 \sin \theta_1 = n_2 \sin \theta_2. \quad (2.7)$$

To guarantee total internal reflection (TIR) θ_1 must exceed the critical angle θ_c which can be calculated by setting $\theta_2 = 90^\circ$. TIR means that E_t does not occur and Snell's law simplifies to [13]

$$\sin \theta_c = \frac{n_2}{n_1}. \quad (2.8)$$

2.1.1 Waveguides

Let's consider a planar waveguide composed of three layers like shown in figure 2.2. For the outer layers we impose $n_1 = n_3$. If a beam inside it propagates in such a way that it impacts at the interface at an angle greater than θ_c , TIR will cause it to propagate through the waveguide until it changes its geometry. If θ_c is not reached the wave partially transmits through the outer layers. However, this picture isn't sufficient to describe the propagation of light in a waveguide. A clarification on the derivation process of optical modes is given in the appendix (A.2 Supporting materials). Here we want to focus on the result, which is the existence of quantized optical modes m , depending on materials and geometries. These modes show the electric (and magnetic) field distribution within a waveguide. This is shown in figure 2.3. The base mode $m = 0$ leads to so-called single-mode waveguides. They are unique because they eliminate intermodal dispersion and minimize loss, enabling high-speed, long-distance transmission with superior polarization stability.

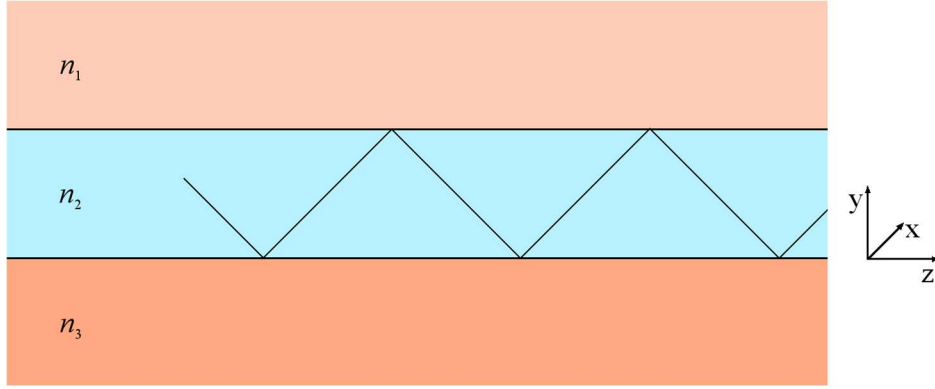


Figure 2.2: Schematic of a planar waveguide composed of three layers. The layers with n_1 and n_3 act as cladding for the "core" with n_2 (taken from [12]).

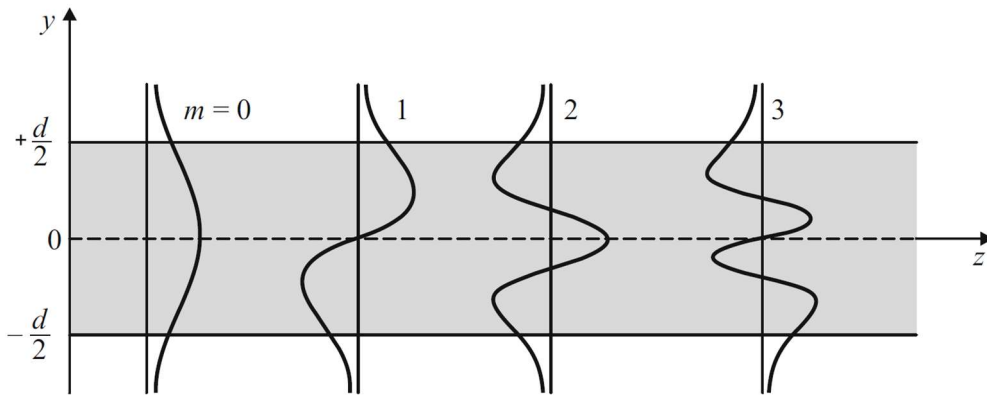


Figure 2.3: Optical modes within a planar waveguide with a core thickness of d (taken from [12] as an adaption from [16]).

In the derivation process the mode depended effective refractive index n_{eff} is discussed. It is the result of the relation (for the z direction)

$$n_{\text{eff}} = \frac{k_z}{k_0} \quad (2.9)$$

of the effective propagation constant in the core k_z to the vacuum one k_0 . n_{eff} links the real phase velocity v to the waveguide geometry and (surrounding) material choice. It is in the range of all n_i 's existing in the waveguide.

There are many variations of waveguide geometries used on SiPh chips like rib waveguides, buried waveguides and inverse rib waveguides [16]. The understanding gained from the planar waveguide can be transferred to all varying geometries. Each waveguide has its own optical modes governed by its geometry and materials. Most of the time given symmetries can be used, to reduce the dimensions of the mathematical problem. The transfer holds also true for optical fibers as another type of waveguide. They are a vital component in most optical (data-) transmission systems. There are many different fiber types, mostly silica (SiO_2) based, with the core and/or the cladding lightly doped to change its refractive index [13]. In the following we will focus on fibers with a (like) step profile. Such a fiber is shown in

figure 2.4 with $n_{\text{core}} > n_{\text{cladding}}$. Note that there are other fiber types such as graded-index fibers with more complex distributions of n [13]. For a more detailed introduction [17] is recommended.

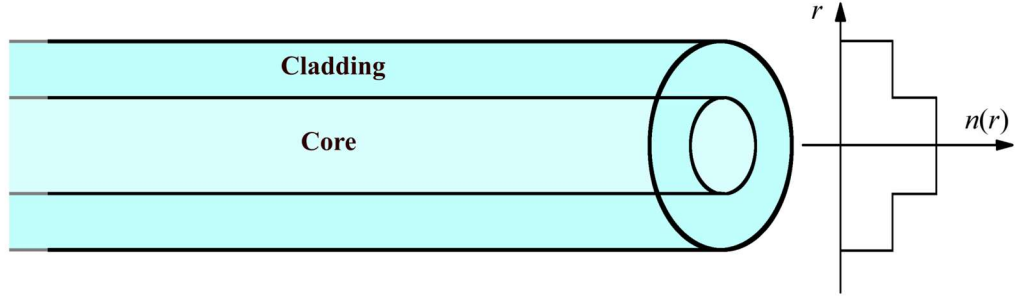


Figure 2.4: Schematic of an optical fiber with a step profile of its refractive index (adapted from [12]).

We are now looking at a cylindrical problem which naturally leads to polar symmetry in the electromagnetic fields' distribution. The boundary conditions need to be met in all directions perpendicular to the propagation direction. This leads to mode fields as shown in figure 2.5. The fundamental mode mathematically would be given by (modified) Bessel functions. A fiber that conducts several modes is called a multi-mode fiber (MMF). Often such fibers have a more complex refractive index profile so that all modes have a similar phase velocity v (in consequence only one n_{eff}) [17].

Reducing the core radius leads to a SMF, where only the fundamental mode is supported. The radius needs to be optimized to suit the intended wavelength of the light going through it. Next to the previously mentioned benefits of single-mode waveguides, SMFs also have low production cost. However, they impose cost-increasing restraints on the light source [17]. Figure 2.5 (right) shows the electrical field distribution of the fundamental mode. The mode field diameter (MFD) is the diameter where the electrical field has a value of E_0/e and, consequently, the intensity at ω_0 is at I_0/e^2 . ω_0 is almost identical to the waist of a Gaussian beam, since the Bessel functions composing the mode field of the fundamental mode only deviates slightly from that of a Gaussian beam [14].

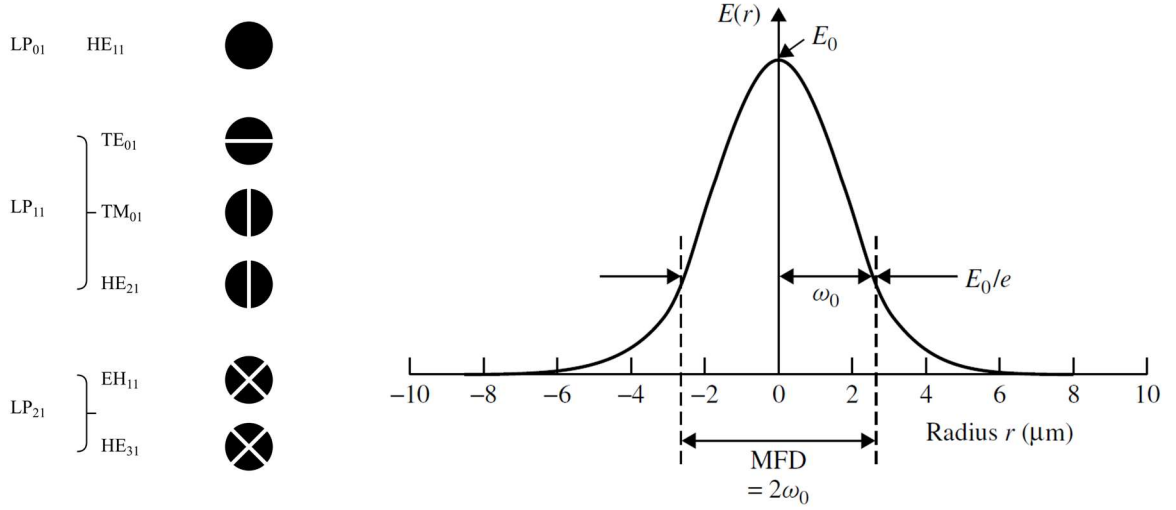


Figure 2.5: Left: Approximated first mode distributions for a fiber with a step profile. The linearly polarized (LP) modes which are superpositions of the exact vector solution. These are classified as transverse electric (TE), transverse magnetic (TM) and hybrid (HE, EH) modes. Right: Mode profile of a SMF (both adapted from [13]).

For some applications, it is paramount to maintain the polarization of the incident light, for which polarization-maintaining (PM) fibers were developed. An example of such a fiber is shown in figure 2.6. The general structure with the core and the cladding remains the same. Additionally, stress elements are introduced in such a way that they create birefringence with a fast and slow axis, referring to the differences in phase velocity by the stress induced refractive index modification along those axes. This suppresses the rotational changes in the polarization plane [17].

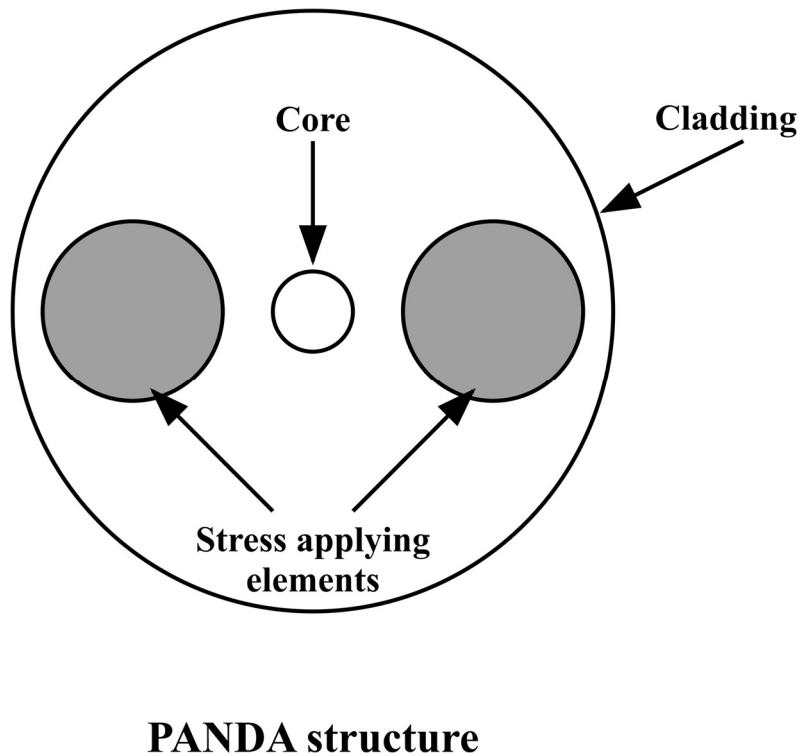


Figure 2.6: Cross-section of a PANDA type PM fiber (adapted from [17]).

2.1.2 Gaussian beams

So far, we concentrated on guided beams. We established that the fundamental mode basically has a Gaussian profile. Now we want to see what happens at a flat fiber to free space interface. A monochrome beam inside a SMF and just after the interface the beam is spatially coherent. Hence, a fixed phase relationship exists across the beam cross-section [18]. Due to diffraction, the beam diverges according to the equation [14]

$$\omega(z) = \omega_0 \left[1 + \left(\frac{\lambda z}{\pi \omega_0^2} \right)^2 \right]^{1/2} = \omega_0 \left[1 + \left(\frac{z}{z_R} \right)^2 \right]^{1/2} \quad (2.10)$$

where we fixed the propagation direction to the z -axis and set the fiber-vacuum interface to $z = 0$. ω is the waist radius, ω_0 the waist radius at the interface and inside the fiber, λ the wavelength of the beam and z_R the Rayleigh length. The latter defines the z value where the waist reaches $\sqrt{2}\omega_0$ and from there on, ω follows a linear increase with regard to z . It should also be noted that the light after the interface has curved wavefronts opposed to the flat ones inside of the fiber. In most cases the divergence angle Θ is calculated by $\Theta \approx \omega_0 \cdot z_R^{-1}$.

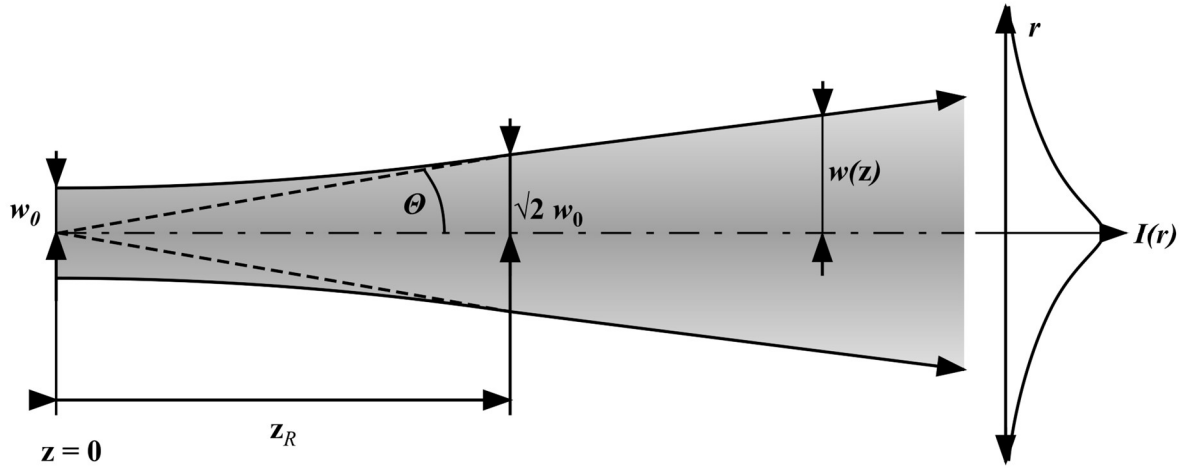


Figure 2.7: Schematic of a Gaussian beam outside a fiber-vacuum interface at $z = 0$ (adapted from [14]).

2.1.3 Grating couplers

To build a working data transmission system, fiber-chip coupling is an essential component. This is where we expect the highest optical losses in the system. The most common coupler types to transfer light from an optical glass-fiber to a photonic integrated circuit (PIC) are edge couplers and grating couplers (GCs). Here we will focus on the latter.

GCs enable us to couple a beam of light into the chip plane from above the chip surface. In principle they use a lattice to utilize Bragg refraction. In figure 2.8 we can see the waveguide is realized within the upper Si layer between the upper cladding and the SiO₂ layer. The lattice is composed of a teeth-like structure within the Si layer with a period of Λ . The plateaus of the lattice have a width of w and the valleys a depth of d . Like with normal Bragg refraction this lattice has diffraction orders for incoming (and outgoing) light at different angles Θ above the waveguide plane. To couple efficiently, the -1st order is taken, since the 0th is not refracted and provides the highest power transmission, while the phase matching condition is still met. This condition dictates that the propagating wave inside the waveguide with n_{eff} , let's say along the z-axis, and the z-axis of the beam within the cladding with n_c must have an identical phase. The phase matching condition can be formulated as

$$k_{\text{waveguide}} = k_{z,\text{cladding}} = k_0 n_c \sin \Theta . \quad (2.11)$$

In subsection 2.1.1 we discussed that for the planar waveguide we have a higher refractive index inside the core with respect to the surrounding material. For the situation at hand this means n_{Si} (or n_{eff}) $> n_c$ where n_c refers to the cladding. The propagation in the waveguide is given by $k_{\text{waveguide}} = k_0 n_{\text{eff}}$. Since the $n_{\text{eff}} > n_c$ and $\sin \Theta \leq 1$ this implies that $k_{\text{waveguide}} > k_{z,\text{cladding}}$. Therefore, direct phase matching is impossible using a classical refractive interface (like a mirror), as momentum cannot be conserved. To overcome this mismatch, the grating introduces a reciprocal lattice vector $K = 2\pi/\Lambda$. The new phase matching condition for the diffraction order p becomes

$$k_{\text{waveguide}} + \frac{2\pi p}{\Lambda} = k_0 n_c \sin \Theta . \quad (2.12)$$

p is an integer counting the modes. $k_{z,p}$ is the corrected propagation for each mode, which we are trying to match to the propagation constant within the waveguide. Since $k_{\text{waveguide}}$ is too big, only negative p values have a physical result, usually $p = -1$ is targeted as mentioned before. This leads to

$$\Lambda = \frac{\lambda_0}{n_{\text{eff}} - n_c \sin(\Theta)} , \quad (2.13)$$

where λ_0 is the vacuum wavelength and n_{eff} the effective refractive index of the coupler region, where the other design parameter d is implied [19].

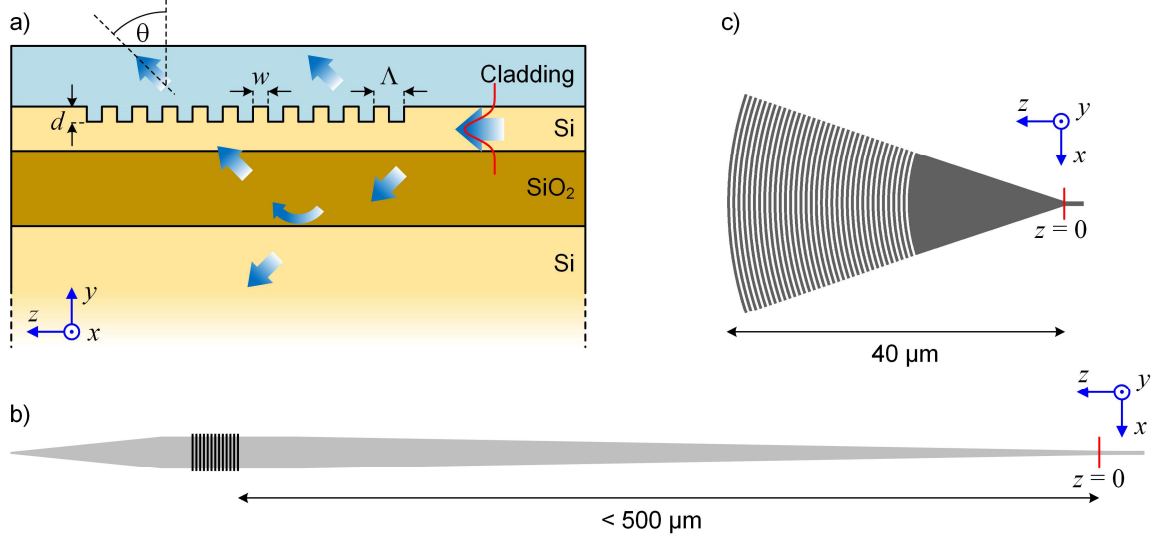


Figure 2.8: 2D schematics of grating couplers. a) Through-chip cross-section of a grating coupler lattice over a waveguide. b) Top view of a conventional grating coupler with a linear adiabatic taper. c) Top view of a focusing grating coupler (taken from [3]).

Real GCs are designed in such a way that they adapt their fields to easily couple into SMFs. Therefore, they need to be of a Gaussian shape with a waist of the mode field within the fiber. Taking the axis from figure 2.8 the z direction is shaped by modification to the lattice constants. To widen the beam mostly geometrical measures within the y -plane are taken like adiabatic tapers or focusing GCs with lens-like shapes.

Effectively, in the coupling process the incoming field excites the possible modes of the receiving structure. The efficiency Γ is given by the overlap of two fields after their respective structure-vacuum interface, which is identical to the square root of the ratio of received to incoming power. The formula is given by [13]

$$\Gamma^2 = \frac{\left[\int_{-\infty}^{\infty} dy \int_{-\infty}^{\infty} E_a E_b dx \right]^2}{\int_{-\infty}^{\infty} dy \int_{-\infty}^{\infty} E_a^2 dx \cdot \int_{-\infty}^{\infty} dy \int_{-\infty}^{\infty} E_b^2 dx} . \quad (2.14)$$

E_a and E_b are the involved mode fields. Γ is influenced by the matching of the MFDs and the distance, placement and angular alignment between the coupling partners. Setups with SMF-GC-coupling have typical insertion losses of 3 dB i.e. $\Gamma^2 = 50\%$ [20]. Highly optimized GC designs can further decrease the losses to 3 dB i.e. $\Gamma^2 = 89\%$ [21]. The resulting spectral shape of two GCs connected via a waveguide is shown in figure 2.9. The coupling can only be optimized for a single wavelength, here 1550 nm. The attenuation of the coupling process itself was omitted.

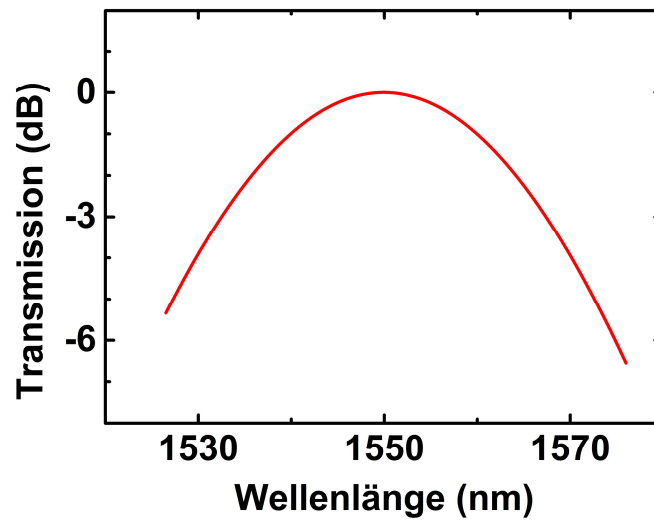


Figure 2.9: Typical spectral shape of a waveguide coupled via an input and output grating coupler (taken from [12]).

2.1.4 Electro-optical modulators

There are several optical modulation mechanisms. They all entail a change in the optical field via an applied (electrical) signal. First, the Pockels effect (linear electro-optic effect): An applied electric field induces a displacement of ions or charge carriers within certain crystals, such like lithium niobate (LiNbO_3). This results in a linear change in the refractive index that is directly proportional to the electric field strength. Due to the centrosymmetric nature of the Si crystal, this effect is absent in SiPh modulators. Next the Kerr effect: While present in Si, it is relatively weak. In this phenomenon, an external electric field induces a third-order nonlinear polarization of the electron shells of the silicon atoms. This distortion of the electron distribution modifies the material's permittivity, thereby changing the phase velocity of the optical beam [22]. Then the Franz-Keldysh effect: Applying voltages to a semiconductor tilts its band structure. This enables photon-assisted tunneling therefore increasing absorption. The modulation is done by direct intensity reduction of the beam. This mechanism is negligible for most optical bands in Si. Next, the thermo-optic effect: It is composed of two competing mechanisms. The first is the thermal expansion of the lattice, which usually contributes less to the overall change. The second and dominant mechanism is the temperature-induced reduction of the band gap, which increases the polarizability [13] and consequently the refractive index. While effective, the thermo-optic effect is comparatively slow.

However, in the framework of the SOI technology and high-speed data transmission the plasma dispersion effect (PDE) is the widely adopted modulation principle. It is also called free carrier effect and works by carrier injection or depletion. Given a rib waveguide with integrated pn-junction like shown in figure 2.12, the application of an external field injects and drains charges into or out of the depletion region. The depletion region and the rib fall together. Effectively, there the charges act as a free charge carrier gas and insulate the electromagnetic field of the beam from the background. The change in charge carrier density leads to the modulation of the effective refractive index. This change Δn is given by

$$\Delta n = \frac{-e^2 \lambda_0^2}{8\pi^2 c^2 \epsilon_0 n} \left(\frac{N_e}{m_e^*} + \frac{N_h}{m_h^*} \right). \quad (2.15)$$

The indices distinguish between electrons e and holes h . N are the respective free charge carrier concentrations and m^* their effective masses. For fast modulation a reverse bias (RVB) is used. RVB means that a voltage in blocking direction of the pn-junction is applied. This drains the depletion region. This is fast since it simply widens the depletion region. No current across the pn-junction is generated. So, it is a pure field effect limited by capacitance and resistance (“RC bandwidth limitation”). The DC-RVB is then modulated with a high-speed radio frequency (RF) signal, enabling modulation frequencies of 50 GHz and beyond [15].

If the pn-junction were to be modulated in a DC forward bias (FWB) i.e. a current over the pn-junction exists, speeds would be comparatively limited. This limitation stems from the minority charge carriers injected into their respective regions. There they need to recombine if the voltage across the pn-junction has decreased or flipped during modulation, which is a comparatively slow process.

Through the carrier injection and depletion also the absorption is modulated. This $\Delta\alpha$ is given by the Drude-Lorenz equation [23]

$$\Delta\alpha = \frac{e^3 \lambda_0^2}{4\pi^2 c^3 \epsilon_0 n} \left[\frac{N_e}{\mu_e (m_e^*)^2} + \frac{N_h}{\mu_h (m_h^*)^2} \right] \quad (2.16)$$

and mostly negligible in its magnitude. Here μ is the mobility of either electrons or holes (denoted by the index) and e the elementary charge

Therefore, in the following it is implied that the change in the refractive index, and the thus resulting change of phase velocity of the light, is the mechanism that drives the modulation in our modulators.

The first modulator type we will look at closer are Mach-Zehnder modulators (MZMs). In principle they are Mach-Zehnder interferometers. In the context of a SiPh SOI device they consist of a beam splitter followed by two independent waveguides, which have the function of the arms of the interferometer. Then the two arms merge in a combiner. So, an initial beam is split and guided through the arms as two separate (in most cases equal) parts. The arms may vary in any parameter, generating a phase shift between the two beam halves. In silicon photonics (SiPh) MZMs modulate by putting either one or both arms within a pn-junction. There we use the PDE to achieve modulation. As introduced with the PDE, we effectively change the phase, therefore also change the phase relation between the two arms. Unlike classical interferometers, a real path length difference is not needed. However, a path difference might be used to set a predetermined working point. After this modulation stage the beam halves are recombined interfering either constructively or destructively.

Figure 2.10 a) shows a rudimentary sketch of a MZM indicating the two different phases within the two arms. Figure 2.10 b) shows a typical electric field and intensity spectrum ($I \propto E_{\text{out}}^2$) of a symmetrical MZM, where the latter acts like $\propto \cos^2(\varphi_1 - \varphi_2/2)$. The phase difference (for an MZM with symmetrical arms) can be calculated by [12,23]

$$\varphi_1 - \varphi_2 = k_1 L_1 - k_2 L_2, \quad (2.17)$$

where L is the length of the individual arm. The indices denote the exact arm with its respective propagation constant k . We recall from subsection 2.1.1 that k is directly linked to the effective refractive index n_{eff} and therefore the phase. To operate the MZM most effectively, the working point is set to the strongest slope point of the intensity curve at $\pi/2$ phase difference. This is the so-called Quad point. The unmodulated phase offset can be achieved by different means like asymmetric DC-RVBs, heaters or an offset between the arms lengths. By modulating the phase relation, we switch between intensity levels on the slope around the working point. To boost the modulation strength, i.e. the intensity difference between highest and lowest intensity, an MZM can be designed with a push-pull configuration in mind. Both arms lead through a pn-junction, which by themselves are connected in series. They are both reverse biased separately while sharing the potential of their series connection point. The modulation signal is applied to this center potential. The effective refractive index in one arm decreases and simultaneously increases in the other. This pushes the distance between the highest and lowest intensity apart during the modulation or synonymously increases the phase differences.

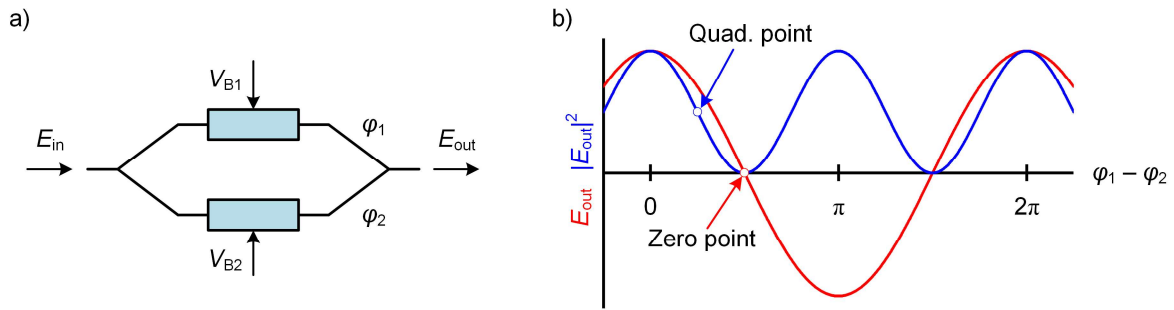


Figure 2.10: a) Schematic of a Mach-Zehnder modulator, where both arms can be modulated via an external voltage. The modulation results in a phase difference between the two arms. b) Resulting electric and squared electric field (the latter is proportional to the intensity) after recombination of the beams over their relative phase shift. The Quad point is the desired center point of the phase modulation for a maximal signal (taken from [3]).

The other modulator type we will discuss are ring modulators (RMs). A simplified sketch of such a modulator is shown in figure 2.11. It consists of a bus waveguide next to a circular waveguide, sometimes called a resonator or cavity. They are not connected but loosely coupled through their proximity. In normal conditions (bottom of the figure) the beam propagates through the bus waveguide, without being altered. If the resonance condition

$$m \lambda_{\text{res}} = 2 \pi r \cdot n_{\text{eff}} \quad (2.18)$$

(and a certain coupling condition) is met, the intensity at the end of the bus waveguide decreases significantly. r is the rings radius, n_{eff} the effective refractive index within the ring, λ_{res} the resonant vacuum wavelength and m a positive integer. Intuitively we can assess the resonance condition, which is that we count the number of full wavelengths fitting into the ring. The distance between the bus and the ring is small enough, so that an evanescent field couples into the ring. At a resonance wavelength a strong stable field forms within the ring by interference. The coupling strength of the ring depends on several factors, including the distance between the bus and ring, coupling length, mode confinement and phase matching [19]. At resonance, a strong field builds up within the ring and couples back into the bus waveguide, where it interferes destructively with the transmitted light. The resulting output intensity spectrum is mostly flat, featuring deep notches at resonant wavelengths as shown in figure 2.11 (right).

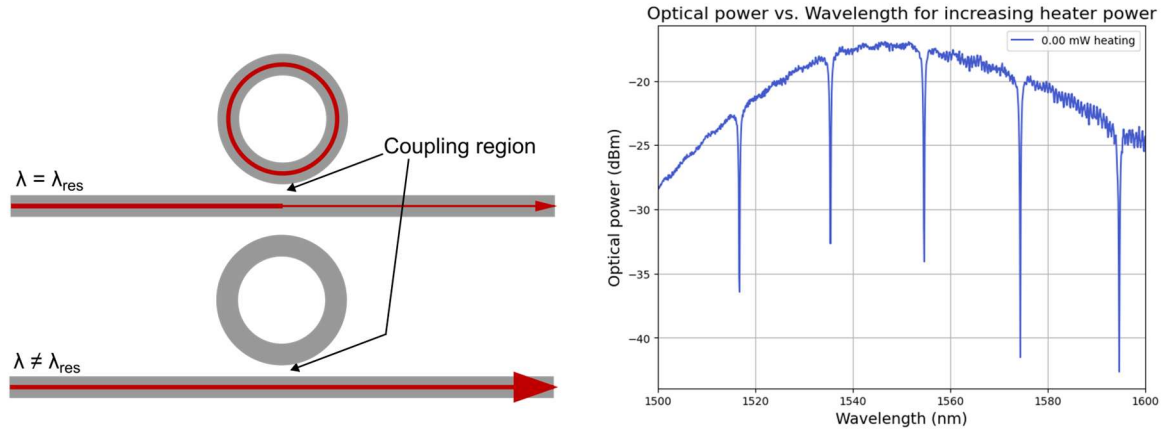


Figure 2.11: Left: Sketch of a ring modulator within resonance (top) or outside resonance (bottom). Right: Spectrum of a ring modulator optically connected via grating couplers.

However, the depth of these notches is strictly governed by the coupling condition. To achieve maximum signal extinction, the device should ideally operate at critical coupling. This condition is met when the self-coupling coefficient t (representing the fraction of light remaining in the bus) and the round-trip attenuation factor a (representing the losses within the ring) are equal:

$$a = t \quad (2.19)$$

Depending on this balance, the resonator operates in three distinct regimes. The first is when $a > t$: Internal losses exceed the coupling strength, resulting in shallow notches (“under-coupled”). Second is the critically coupled situation at $a = t$: Coupling and losses are perfectly balanced, leading to near-zero transmission at resonance. Finally, the “over-coupled” regime with $a < t$: Coupling is stronger than the internal losses, which broadens the resonance and reduces the extinction ratio.

The distance between these notches is called free spectral range (FSR) and is not constant. In the specific measurement shown in figure 2.11, the overall envelope of the spectrum is shaped by the transmission characteristics of the GCs used for fiber-to-chip coupling, as previously discussed in subsection 2.1.3 and shown in figure 2.9.

The pn-junction for RM sits within the ring. Some designs omit the pn-junction from the coupling region. Again, we modulate the RM over n_{eff} via the PDE. The working point is set on the slope either on the rising or falling slope of one notch via three mechanisms outside of the physical design. One pathway is via a DC-RVB, another is by an optional heater layer above the ring utilizing the thermo-optic effect. The final one is the wavelength, which is used for our experiments. Like with a MZM the application of an RF electrical signal to the pn-junction shifts the intensity to higher and lower levels surrounding the working point. Technically the spectrum is shifted by the voltage change of the pn-junction via the resonance condition from above, while the wavelength stays fixed, yielding an output intensity oscillation.

Modern highspeed EO-SOI modulators are based on pn-junctions and utilize the PDE. Nowadays RMs are the favored technology. Nonetheless we will look at both types of modulators. They use cross-sections like shown in figure 2.12. Production begins with a pre-prepared SOI wafer composed of a Si substrate layer, a buried oxide (BOX) layer and a thin layer of monocrystalline Si. First, all different

dopings are created by ion implantation. In the next step the slab and the rib are etched out of that thin Si layer. The height offset between the rib and the slab is called etch depth. The rib is the waveguide, whereas the slabs will act as electrical contacts to the pn-junction within the rib. The slab is doped similar to the pn-junction, however at much higher doping levels for improved electrical conductivity. This is indicated via the plus-signs. After a high temperature healing step, a SiO_2 passivation layer is added. It also serves as the cladding of the rib-waveguide. In an intermediate step a heater might be added via metal deposition. Finally, paths for the metal contacts are etched and filled with metal. These serve as electrical contacts to the pn-junction [19].

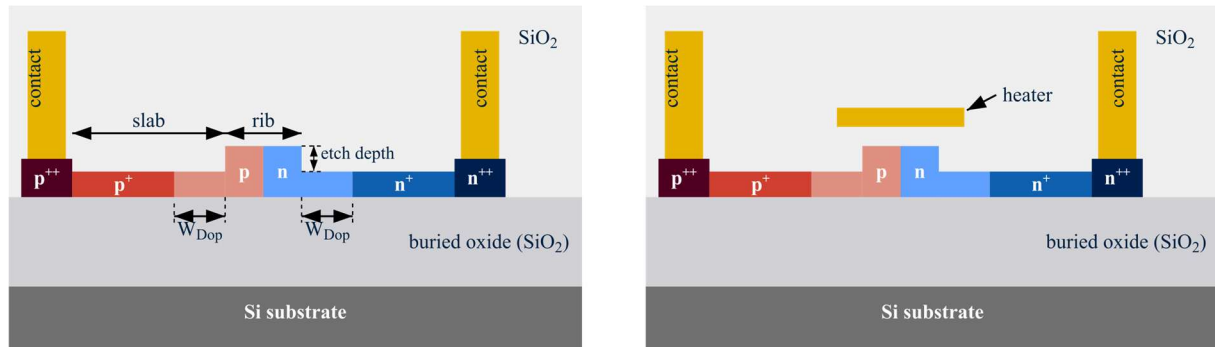


Figure 2.12: Cross-sections of electro-optic modulators. Left: Demonstration of the most important design dimensions and relative doping concentrations as used in MZMs. Right: The same cross-section but with a heater over the rib waveguide/pn-junction, used in RMs (adapted from [24]).

2.2 Irradiation damage, hardening and annealing mechanisms in Silicon

To confirm the viability of the application of SiPh modulators in harsh X-ray radiation rich detector environments, the irradiation damage mechanisms and consequences need to be investigated. Since this is a developing field, not many publications in book form are available, however [25] delivers a valuable overview. As a standard resource [26] is recommended.

Radiation in high energy physics (HEP) experiments is differentiated in charged and charge-neutral radiation. The latter is mainly comprised of neutrons and X-rays. Their energy usually is in the MeV range. Received radiation dose is split by the exchange mechanism between radiation particles and irradiated matter. In the case of X-rays, it is designated by the total ionizing dose (TID), which accounts for total received energy through irradiation. It is counted in Gray: $\text{Gy} = 1 \text{ J/kg}$. In case of the most inner detector layer of the HL-LHC upgrade the expected TID is in the range of 11 MGy [27,28] (the Technical Design Report for CMS [29] states a TID of 12 MGy – the one for ATLAS states 10 MGy [30]). As TID implies, the exchange mechanism is ionization, either through the photoelectric effect or by Compton scattering. In either case an electron-hole-pair is created, leading to excess charge. By themselves they already alter the optical and electrical properties of the device via the PDE and free carrier absorption. In RVB due to their larger mobility, the now free electrons get drained and leave behind the holes, creating a positive electric field. Additionally, the holes may be trapped within the SiO_2 or the Si- SiO_2 interface, creating a semi-permanent electric field [24]. For modulators the latter is the main concern. At the interface the lattice is perturbed forming strong traps, where the holes get captured. Therefore, the positive charge distribution at the interface is exceptionally high with higher

TIDs, since additional holes travelling through the SiO_2 also get trapped there. At the p-slab for any given modulator, this results in a so-called pinch-off (figure 2.13), which may counteract externally applied voltages. This leads to a collapse in the modulation capacity of the pn-junction. The pinch-off is one-sided since the conduction is dominated by the majority charge carriers within a doped region. On the p-side the holes are repelled from travelling through the pinch-off region by the positive electric field of the trapped charges. On the n-side there is an accumulation of electrons which does not interfere with conduction. The pinch-off is the primary failing mechanism.

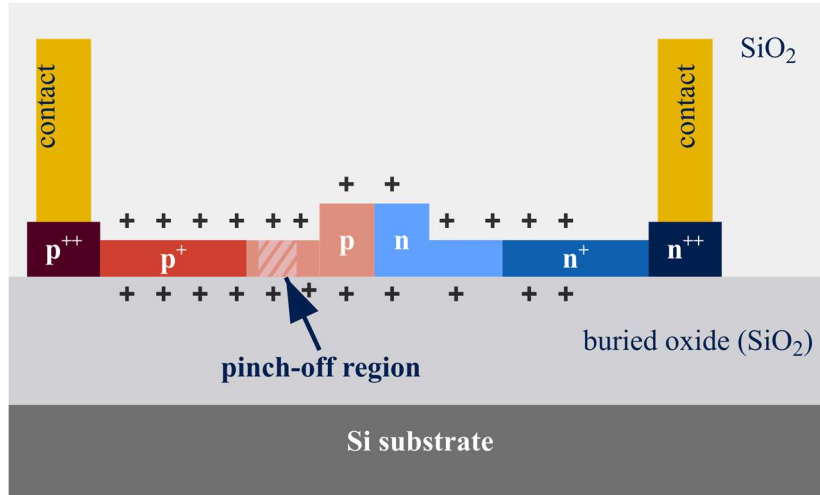


Figure 2.13: Pinch-off in the p-doped slab region (adapted from [24]).

The charge entrapment at the interface is part of the surface damage. The other kind of surface damage also occurs at interfaces, where Si-H passivizes otherwise free valence electrons. X-rays destroy that bond and leave so-called dangling bonds. Due to the abundance of holes, the now unpaired electrons recombine with the abundant holes, leading to error currents.

The modulators will also face strong neutron irradiation, which will not be investigated in this thesis. To evaluate SiPh modulators further, independent neutron irradiation studies must be conducted. The primary damage mechanism in that case is the displacement of lattice atoms through their kinetics. So, one or several atoms are removed from the lattice [28,31]. Defects of this nature fall under the denomination of bulk damage. However, the effects of neutron irradiation on modulators are most likely negligible compared to X-ray irradiation.

We seek to mitigate the consequences of the radiation damage. There are three general approaches to harden the modulator designs against radiation, therefore shifting the exposure limit to higher TIDs. Reference [32] presents them in the modulator context:

1. Etching depth of the slab: Increasing the thickness of the slab physically increases the distance between the trapped charges surrounding the slab. Therefore, a higher concentration, i.e. a higher field, is necessary to achieve the pinch-off. This comes at the cost of a lower confinement of the optical mode within the rib potentially leading to higher optical losses.
2. Increasing the doping of the slab counteracts the electric field of the trapped charges, again increasing the TID to pinch-off. But higher doping means increased absorption through the surplus of free charge carriers. Furthermore, the capacitance of the structure increases, decreasing its bandwidth limit.

3. Complex doping profiles on the slab towards the rib. This is a more surgical approach compared to simply lifting the doping levels, however this comes at the cost of further complicating the designs [24].

In practice a balance between the first and the second or all three strategies are applied simultaneously. Another part of the irradiation mitigation strategy is to utilize annealing. In this context, annealing refers to the recovery of the modulation capabilities via two methods: thermal annealing, which occurs over time at ambient or elevated temperatures, and current injection annealing by the application of a FWB. The former follows Arrhenius's behavior. The rate of the trapped-hole-release depends on the applied temperature. Since SiPh devices are not universally resilient to higher temperatures, this annealing method is not time efficient [32-34].

Therefore, FWB annealing is the favored method. FWB means we apply voltages in such a way that the pn-junction is conducting. Minority charge carriers, i.e. the electrons, are injected into the p-region, which leads to their increased density, in turn increasing the chance of a tunneling event. Originally, two hypotheses were introduced to discuss what happens next on a microscopic level. The first one suggests the electrons tunnel through the Si p-slab towards the trapped holes. The other says holes tunnel towards the Si valence band. It was shown that this tunneling successfully heals the pinch-off [26].

2.3 High frequency electronics

The surrounding electronics of single EO-modulators partially need to run at the same speed as the target modulation speed. In the gigahertz regime we can no longer just work with voltages and currents, but we need to take field effects into account. Radio Frequency (RF) electric conductors and components are their own complete domain. Therefore, we will limit ourselves to the basics, following [35,36]. In the RF domain, just like with optical conductors, we need to take the conductor geometry and materials surrounding the "main" conductor into account. Figure 2.14 shows the cross-section of a coaxial cable. It will serve as an example to transfer the knowledge gained from the theory surrounding it. A coaxial cable consists of an inner conductor, an outer conductor and a dielectric separating the two. If an RF signal is pushed through the cable, an oscillating electric field (red) is generated between them through the dielectric. Therefore, a capacitor-like behavior exists. Additionally, the current in the inner conductor yields an additional magnetic field (green), generating currents. This basically describes an inductor, which therefore resists the change of the magnetic field demanded by the oscillating signal. This can be modeled as a series inductance. In summary, an infinitely small coaxial conductor can be approximated as series of a resistor and inductor (the former is the cables inherent resistance) and a shunt capacitor in parallel with a shunt resistor. The equivalent circuit diagram with a cable length of Δx is also shown in figure 2.14. Note that the shunt resistance $G' \Delta x$ takes leakage currents into account, since a perfect insulator does not exist. It can be shown that the cable acts like two voltage sources at each end. This enables us to calculate the characteristic impedance

$$Z_0 = \sqrt{\frac{i\omega L' + R'}{i\omega C' + G'}} \quad (2.20)$$

This is a critical design parameter.

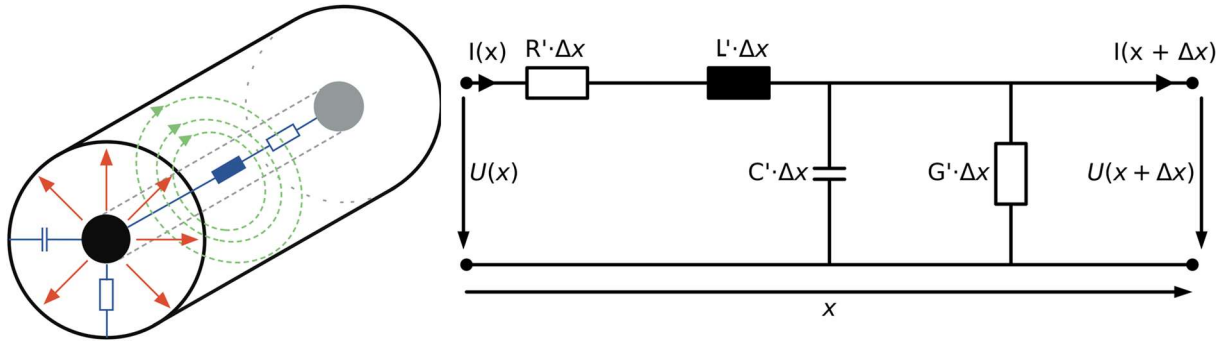


Figure 2.14: Left: Cross-section of a coaxial cable, its fields during conduction (red: electric, green: magnetic) and an indication of the equivalent circuit diagram (blue). Right: Equivalent circuit diagram of an infinitesimal coaxial cable length with the scaling equations (both adapted from [35]).

For any given RF circuit ideally its individual parts are matched to the same characteristic impedance. This process is called impedance matching. For each part an individual equation for Z_0 can be found. If the characteristic impedances are not well matched, reflections are introduced into the circuit, possibly up to the point of total reflection. In the framework of EO-modulators, which also should be matched, that could imply reduced modulation voltages at the pn-junction. To avoid reflections at the end of a circuit, a termination impedance is needed. More specifically a resistance, since an impedance

$$Z = R + iX \quad (2.21)$$

has a resistance R and a reactance X , where only the former dissipates energy preventing reflections. Most common is a characteristic impedance of $Z_0 = 50 \Omega$ for devices and cables.

To evaluate the quality of the impedance matching, S-parameters are commonly used, specifically the S_{11} (see below). In the RF domain current flow and voltages do not offer a practical way to analyze circuits. Figure 2.15 shows a two-port network. This can be any circuit with RF connections at both of its ends. Basic examples would be a coaxial cable or simply a transmission line. a_i are the complex input amplitudes of the waves going towards the circuit (remember that an RF signal can be modeled as two voltage sources at the two ends), b_i are their reflected counterpart. Then the individual S-parameters can be formulated by calculating all four possible fractions b_i/a_j :

- S_{11} : input reflection coefficient with matched output
- S_{12} : reverse transmission coefficient with matched input
- S_{21} : forward transmission coefficient with matched output
- S_{22} : output reflection coefficient with matched input

The a_i not used in each fraction is set to 0, meaning either the input a_1 or the output a_2 has perfect impedance matching. S_{ij} are also complex numbers. A well-matched circuit has an S_{11} and S_{22} close to 0 or $-\infty$ dB respectively and an S_{12} and S_{21} of 1 or 0 dB respectively. The conversion is made via

$$S_{dB} = 20 \cdot \log_{10}(|S|) . \quad (2.22)$$

To translate the S-parameters to a tangible concept, their square magnitude represents the fraction of power transmitted or reflected.

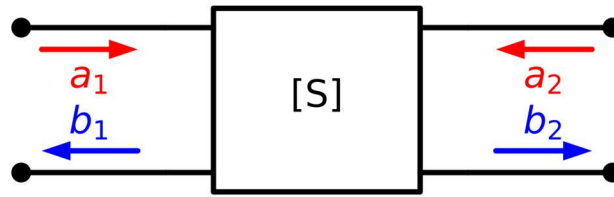


Figure 2.15: Two-port network with associated wave variables a_i and b_i , whose fractions $b_i \cdot a_j^{-1}$ lead to the S-parameters (taken from [35]).

Any conductor in the RF regime, where they are referred to as waveguides, since their behavior is governed by the fields just like for optical waveguides, shows a complex behavior like the coaxial cable. To connect microchips to testing equipment, often bridge PCBs with a cable connector at one end and wire bonds to the microchip at the other are necessary. The resulting circuit and its many components need to be impedance matched, just like described for the coaxial cable example and to quantify the matching the S-parameters ought to be recorded.

Two waveguide cross-sections commonly implemented on PCBs are shown in figure 2.16. On the left side is a so-called coplanar waveguide (CPW), on the right a grounded coplanar waveguide (GCPW). For both the RF signal path is along the center conductor on the PCB surface. The layer thickness of the surface material is T . The center conductor's (the waveguide) width is W . To the “ground planes” (typically they are grounded, hence the name) on both sides of the waveguide a gap of the width G is implemented. Both types as have a dielectric (the PCB material) underneath them. In the case of the CPW design ideally the thickness of the dielectric H goes to infinity (in reality it is simply maximized). For GCPW structures this thickness is another design parameter, since beneath it an additional ground plane is realized. To calculate the exact impedance of such designs is very difficult, since the fields distribution changes in the presence of conductors and dielectrics. The electric field within the dielectric polarizes it via its (relative) permittivity ϵ_r additionally influencing the impedance matching. Therefore, in most cases simulations are used to calculate the characteristic impedance. However, changing the length of a design mostly changes the attenuation of the waveguide. This results in either a steeper (long waveguide) or less steep (short waveguide) S_{12} and S_{21} .

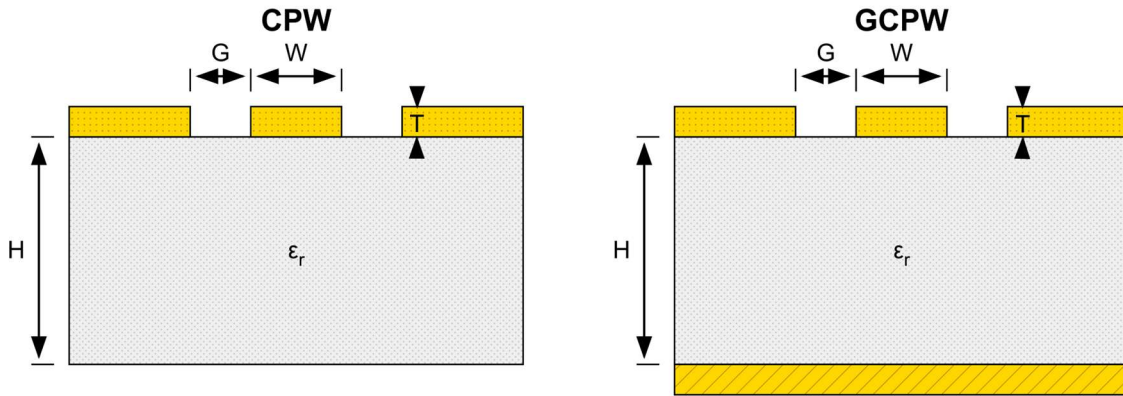


Figure 2.16: Cross-section of a coplanar waveguide (left) and a grounded coplanar waveguide (right).

For the biasing of EO-modulators a vital component of the electrical signal path are the so-called bias-tees. They combine a DC voltage with the RF modulation signal into a single output port. A simple bias-tee is illustrated on the left in figure 2.17. The function is achieved by using a capacitor to block DC from the RF source, while a single inductor isolates the DC supply from the RF path.

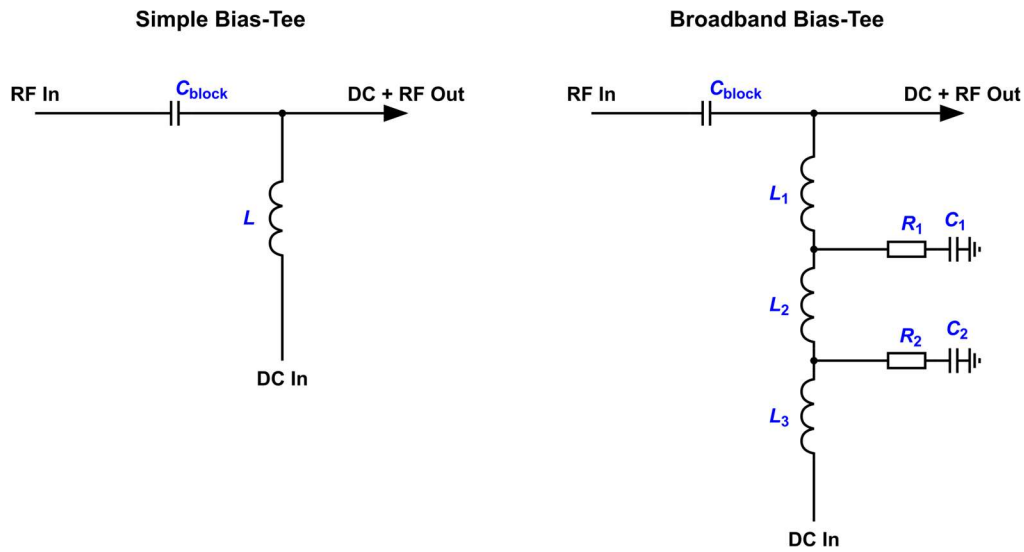


Figure 2.17: A simple (left) and more complex broadband bias-tee (right - adapted from [37]).

However, due to parasitic capacitances, real inductors exhibit a self-resonant frequency (SRF) that limits their effective bandwidth. To cover a wide frequency range, a broadband bias-tee design like in the right of figure 2.17 is required. This involves cascading multiple inductors of decreasing SRF values, where small inductors block high frequencies and larger ones handle the lower spectrum. Since these inductive components typically feature low series resistance, they form a high-Q (high quality factor) network. This means the system stores energy efficiently with minimal dissipation, rendering it highly susceptible to sharp, persistent resonances. To prevent such oscillations, damping is critical. Ferrites are typically integrated to introduce resistive loss, effectively dampening lower frequency oscillations, while RC-shunt networks are employed to “de-Q” the circuit and suppress ringing across both high and low frequencies. Instead of a chain of inductors, also conical inductors may be used, due to their superior bandwidth.

3 The COTTONTAIL chip

In this chapter the chip for detector instrumentation with wavelength division multiplex (COTTONTAIL) will be introduced [7]. A schematic of the chip is provided in figure 3.1. In the following a general introduction to some of the chip's structures is given. The structures used in the experiments in chapter 4 are highlighted. The structures investigated under irradiation in chapter 5 have their own subsection. The was produced using the imec ISIPP50G SOI platform [38]. It is 10.45 mm long, 5.15 mm wide and 0.65 mm tall. The BOX layer is approximately 2 μm thick. The optical Si layer is approximately 211 nm thick and contains all the optical components, like waveguides, modulators and GCs. All designs are optimized for a wavelength of 1550 nm (C-band).

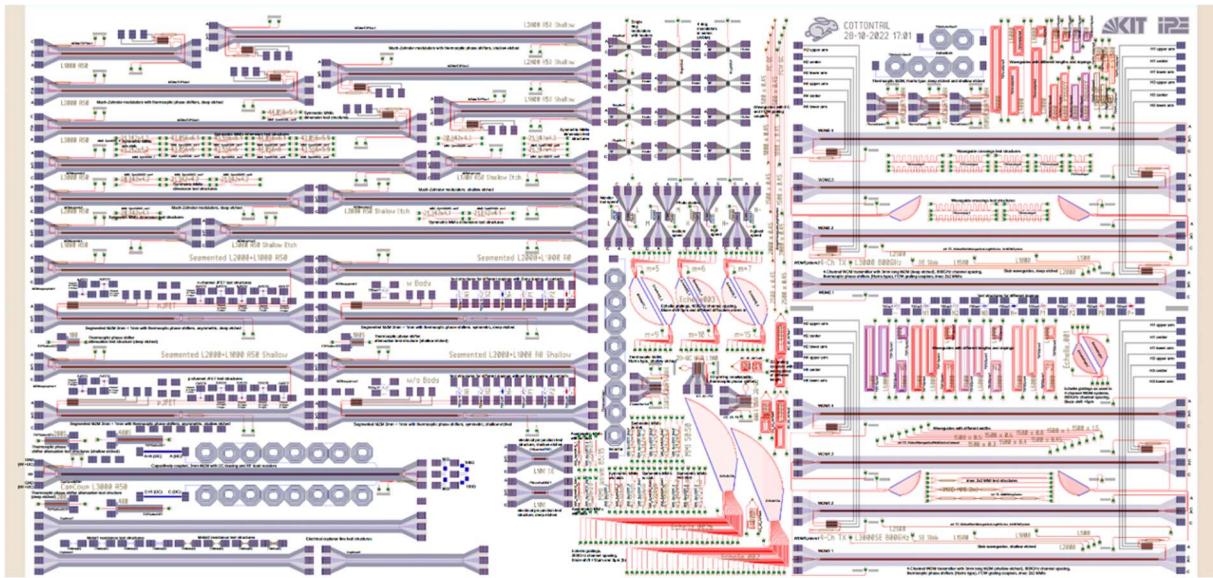


Figure 3.1: The layout of the COTTONTAIL chip (by Marc Schneider).

The chip contains several testing structures apart from modulators. These are in the top right part of figure 3.1, the middle part of the COTTONTAIL chip as shown (rotated) in the top of figure 3.2 and the lower left part shown in the bottom of the same figure. These include photo diodes, isolated multi-mode interferometers (MMI) (as beam splitters), all GCs, isolated Harris-type thermo-optic phase shifters [39] and various kinds of waveguides of different lengths (500 μm to 10 mm), doping concentrations and geometries. Then there are custom structures like electrical inductors, test pn-junctions/MZMs with no optical connections and planar concave gratings also called echelle gratings. The latter are of note, since they are an in-group development started in [4]. A publicly available design program is provided at [40].

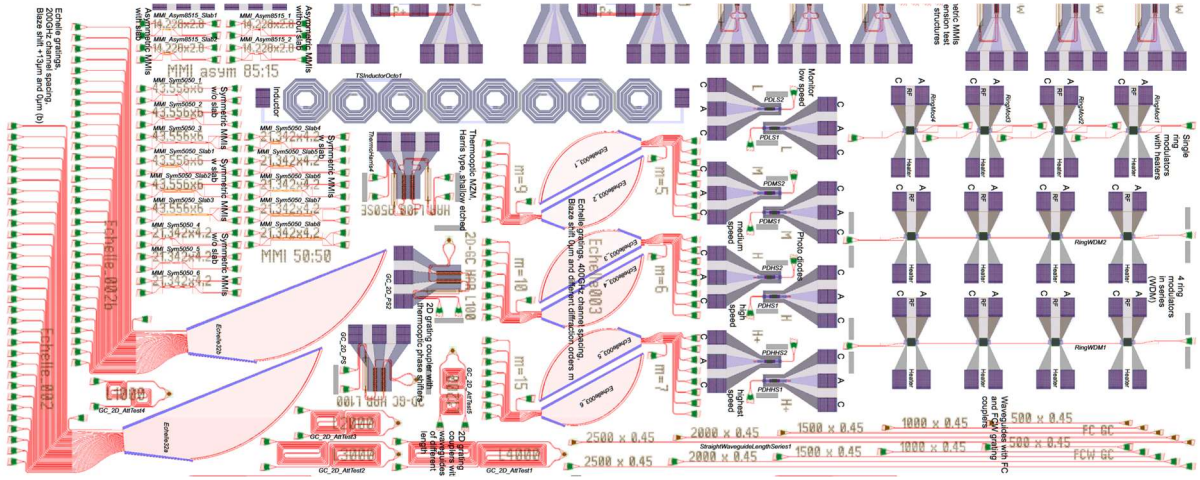


Figure 3.2: Top: Rotated middle section of COTTONTAIL.

Next, on the right side of the chip are two four-channel WDM systems using 3 mm long MZMs for modulation. The two systems differ in the slab thickness of their MZMs i.e. their radiation hardness. Since the GCs are an essential part of each optical structure their previously measured minimum optical loss is 3.57 dB at 1554.5 nm at a coupling angle of approximately 11° .

In section 4.3 the spiral waveguides shown in figure 3.3 are investigated. They differ in their lengths and the waveguide's relative doping concentrations as indicated by their colors. From the straight waveguides on the right bottom in the top part of figure 3.2, the bottom most one is used for optimization measurements in section 4.2 and partially in section 4.3. All waveguides that measurements are performed on use FCW (focusing wideband) GCs.

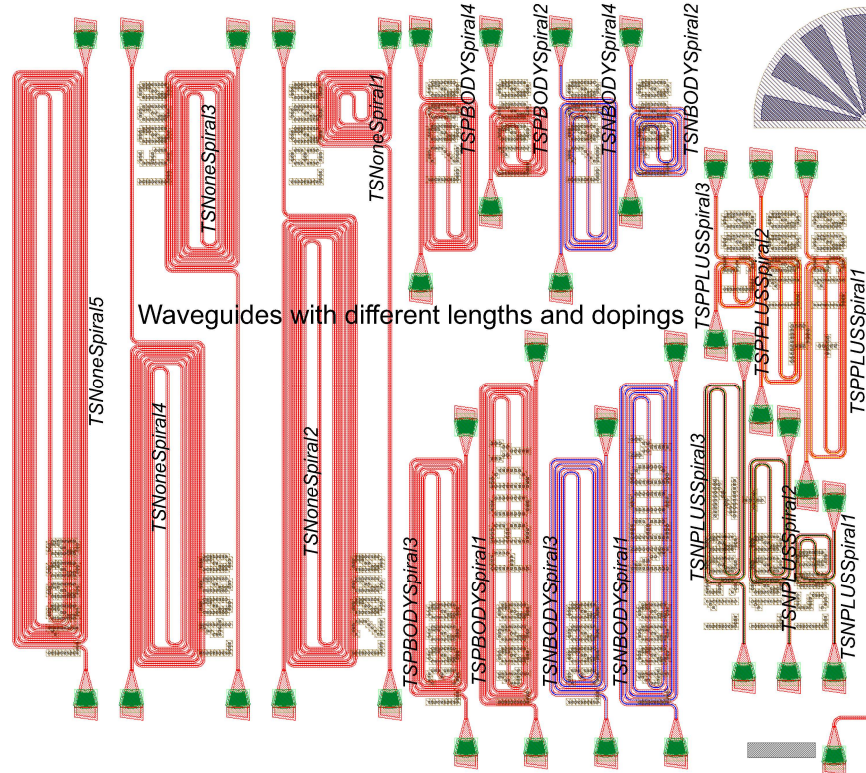


Figure 3.3: Spiral waveguides with different doping levels indicated by their colors.

3.1 Ring modulators

In the top middle part of the COTTONTAIL chip, as shown in the top image of figure 3.2 on the right, three columns of RMs are integrated. The modulators themselves are standard PDK structures. A close-up image of the layouts is provided in figure 3.4. There the left column are single RMs, while the middle and right column consist of systems with four optically series-connected RMs on a single waveguide. We can see two electric connection pads on either side of each RM. The left connectors are for the pn-junction, the right ones for the heater. All pads have a size of $60\text{ }\mu\text{m}$ times $80\text{ }\mu\text{m}$ and a pitch of $100\text{ }\mu\text{m}$. The RMs themselves are impedance matched to $50\text{ }\Omega$, however the pads and their connections to the individual pn-junction are not. The GCs for optical coupling are above and below the individual RM, or system respectively.

The designs parameters like dimensions and doping concentrations are not fully disclosed by imec, as well as ring and coupling region dimensions. Known are $10\text{ }\mu\text{m}$ ring diameter, the 450 nm rib width and 80 nm slab thickness. The ring diameter was determined using the spectral FSR, which is in the region of 19 nm , and confirmed via visual inspection under a microscope. Generally, RMs have comparatively high doping concentrations, which we introduced in section 2.2 as increasing the radiation hardness. The cut-off frequency for the modulation lies beyond 40 GHz . An error free data transmission of up to 11.3 Gb/s could be observed; however, this measurement was limited by the driver electronics. These results are shown in the plots in figure 3.5.

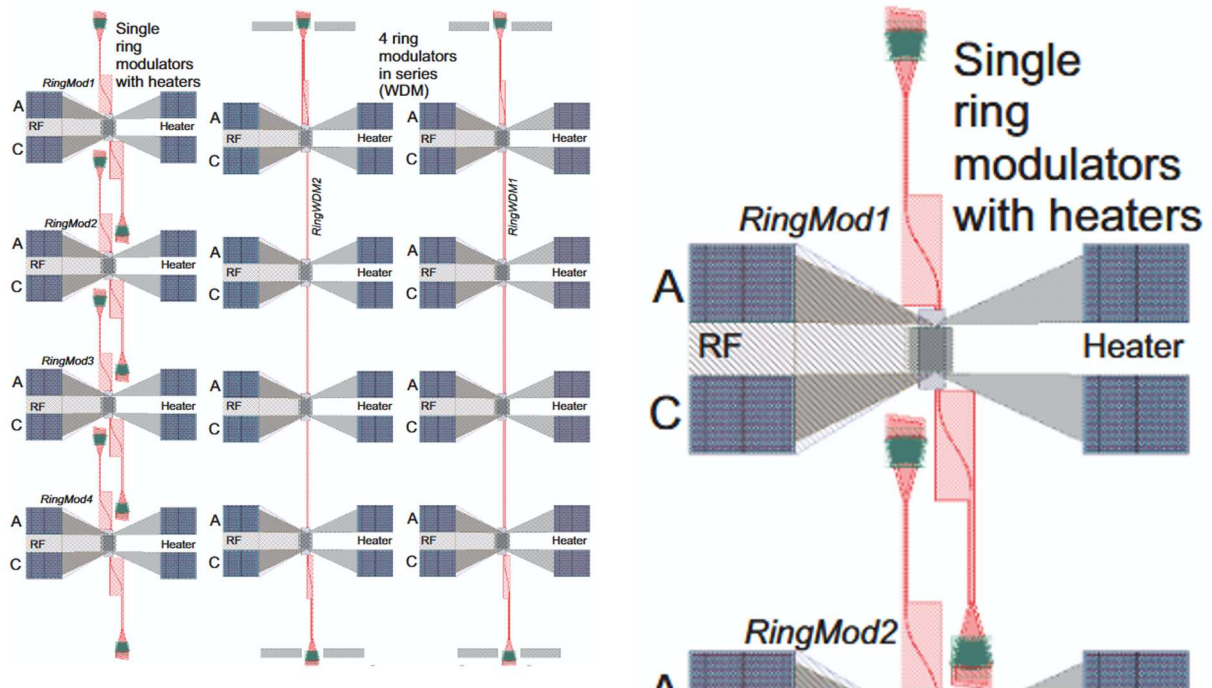


Figure 3.4: Left: Single ring modulators in the left column and two four channel systems to the right. Right: Zoomed in image of the RingMod1 layout.

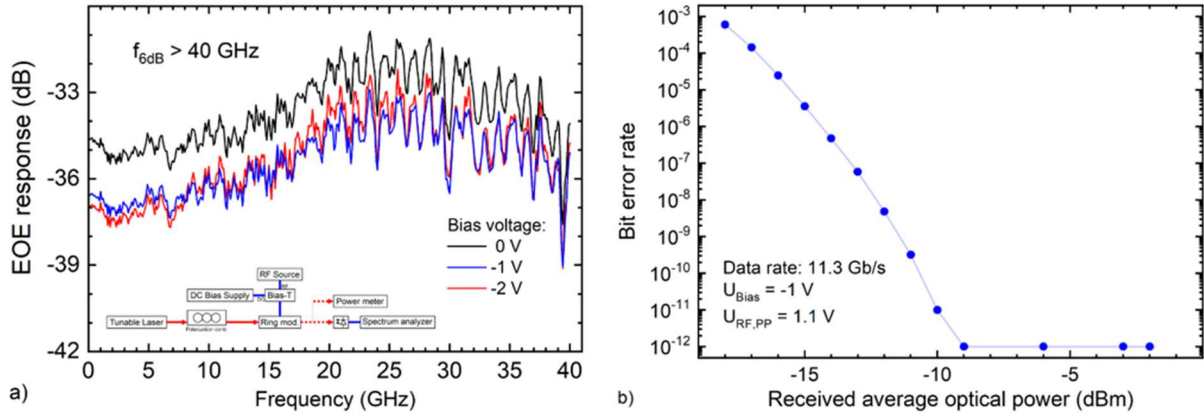


Figure 3.5: a) Modulation power characterization of a ring modulator measurement up to 40 GHz at different bias voltages. b) Bit error rate measurement plot of the same modulator (taken from [7]).

3.2 Mach-Zehnder modulators

The left side of the COTTONTAIL chip is dominated by single MZMs. These include 1 to 3 mm long versions. Of each kind there are radiation hard and standard versions. The former have a slab thickness of 140 nm and are designated by an “SE” (shallow etch) in their name. Standard etch depth MZMs have no additional naming component and a slab thickness of 80 μ m. Of special interest are the MZMs labeled as L1000 A50 Shallow Etch (top of figure 3.6) and L1000 A50 (bottom of figure 3.6). They are investigated in the irradiation measurements discussed in section 5.4. Close to the pn-junction curves in the waveguide arms can be seen. These curves are asymmetrical between the two arms. This is to set an initial phase shift between the arms, which could also be used to set the working point apart from the biasing. Extending from the curves away from the pn-junction both MZMs feature two grating couplers on each end with a MMI in front of them. On the input side the MMI acts as a 50:50 splitter, equally distributing the light between the two arms. On the other side it acts as a combiner. Which GC is chosen on this side is secondary, since it only determines its power relative to the phase shift, with one output being the inverse of the other [18].

A standard etch depth MZM was characterized in [7]. As can be seen in figure 3.7 its 6 dB cut-off frequency was 36.7 GHz, therefore theoretically enabling a bitrate around 40 GHz. Shallow etched 1 mm MZMs showed a roughly 6 GHz lower cut-off frequency.

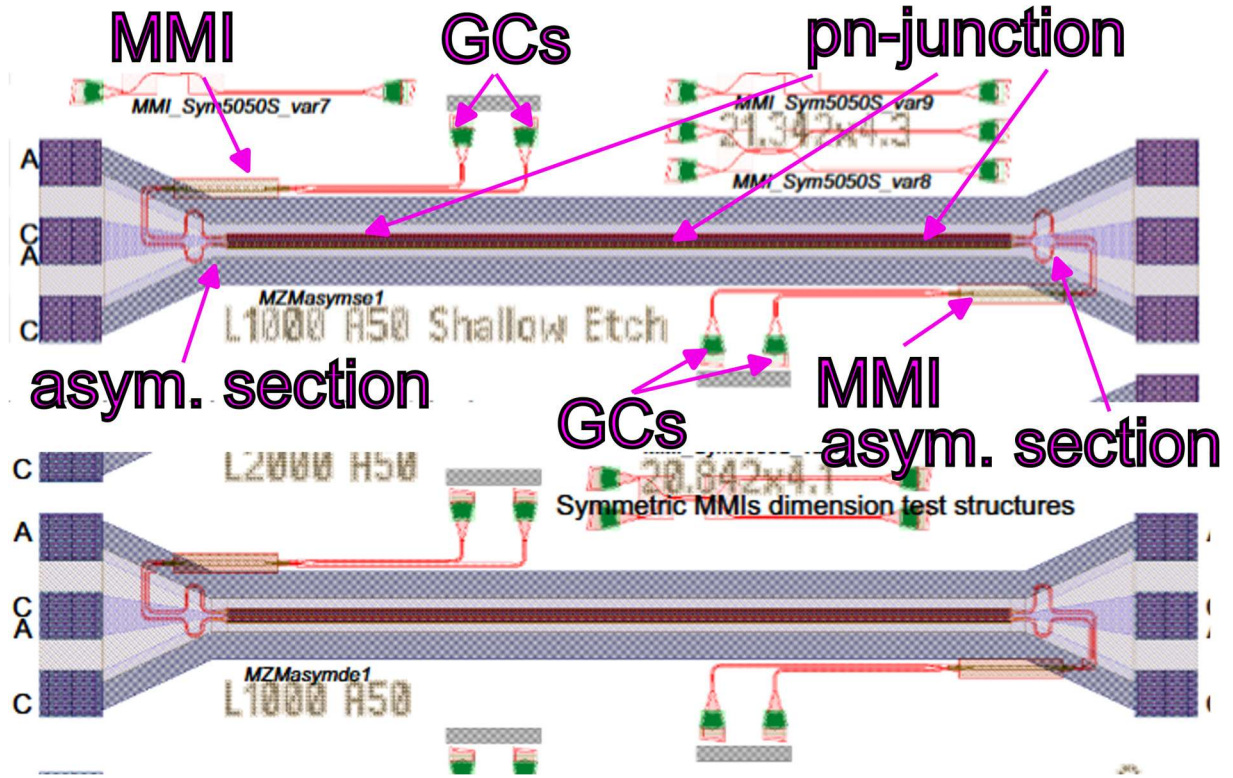


Figure 3.6: Top: Radiation hardened "shallow etch" Mach-Zehnder modulator. Bottom: Standard Mach-Zehnder modulator.

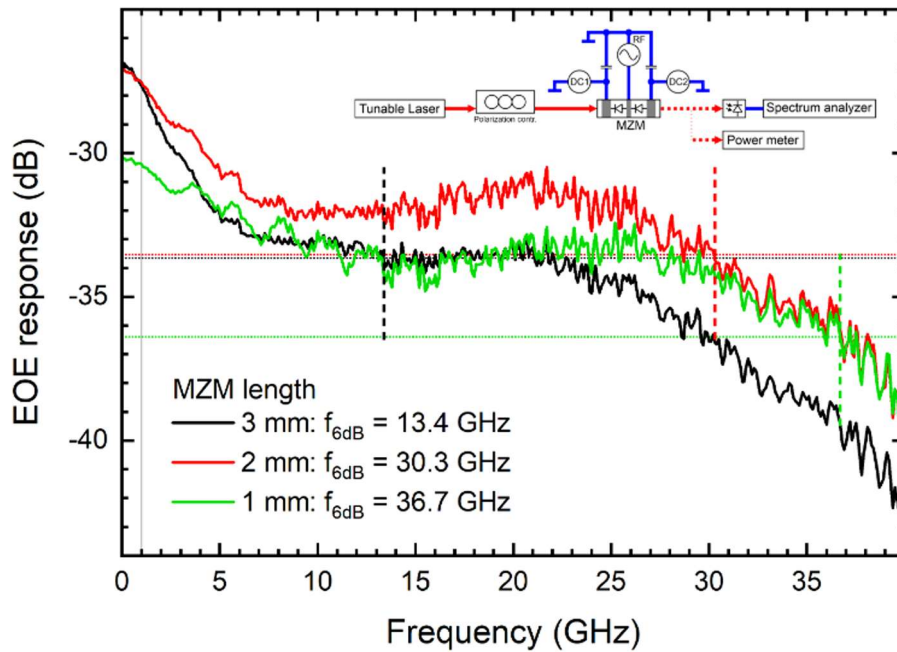


Figure 3.7: Modulation power characterization measurement of different Mach-Zehnder modulators measured up to 40 GHz (taken from [7]).

4 Packaging and interconnection technology

To perform the irradiation measurements of chapter 5 a series of preparatory steps must be performed. The packaging and interconnection technology is part of those preparations are discussed in the following.

4.1 Characterization setup

Figure 4.1 shows the experimental setup used to characterize the COTTONTAIL chip and the fiber-to-chip coupling. It has two distinct configurations, depending on the specific measurement requirements. They are schematically shown in figure 4.2. The DC Configuration is used to quantify the optical yield of a device or structure. The signal chain starts with a tunable laser (Santec TSL-570). The light from the laser is led through a SMF to a polarization controller (Phoenix Photonics), which enables the correct polarization alignment for the next step. This is the optical coupling stage, where an in- and an outbound cleaved SMF with mirror-like facets are positioned at an angle above the GCs of the device/structure. A schematic of this is shown in figure 4.3.

The coupling efficiency of any GC strongly depends on the correct polarization. Since precise alignment in the micrometer range is necessary for efficient fiber-to-chip coupling, each fiber holder is connected to micrometer screw stages for coarse and a piezo stage (Physik Instrumente Nanocube E-564.I3N) for fine alignment. The alignment is optimized in all three dimensions. For stability the chip is secured to the sample stage via vacuum suction, which is applied constantly via a vacuum pump. The fiber holders themselves are custom made with a vernier scale which can be seen in figure 4.1 in the inset. The outbound fiber, which is in any case a standard SMF since polarization is no longer important, leads the light to an optical power meter (OPM). In this case a 81623B Germanium optical power head connected to a 8163B Lightwave Multimeter with an Agilent interface module (all Agilent Technologies).

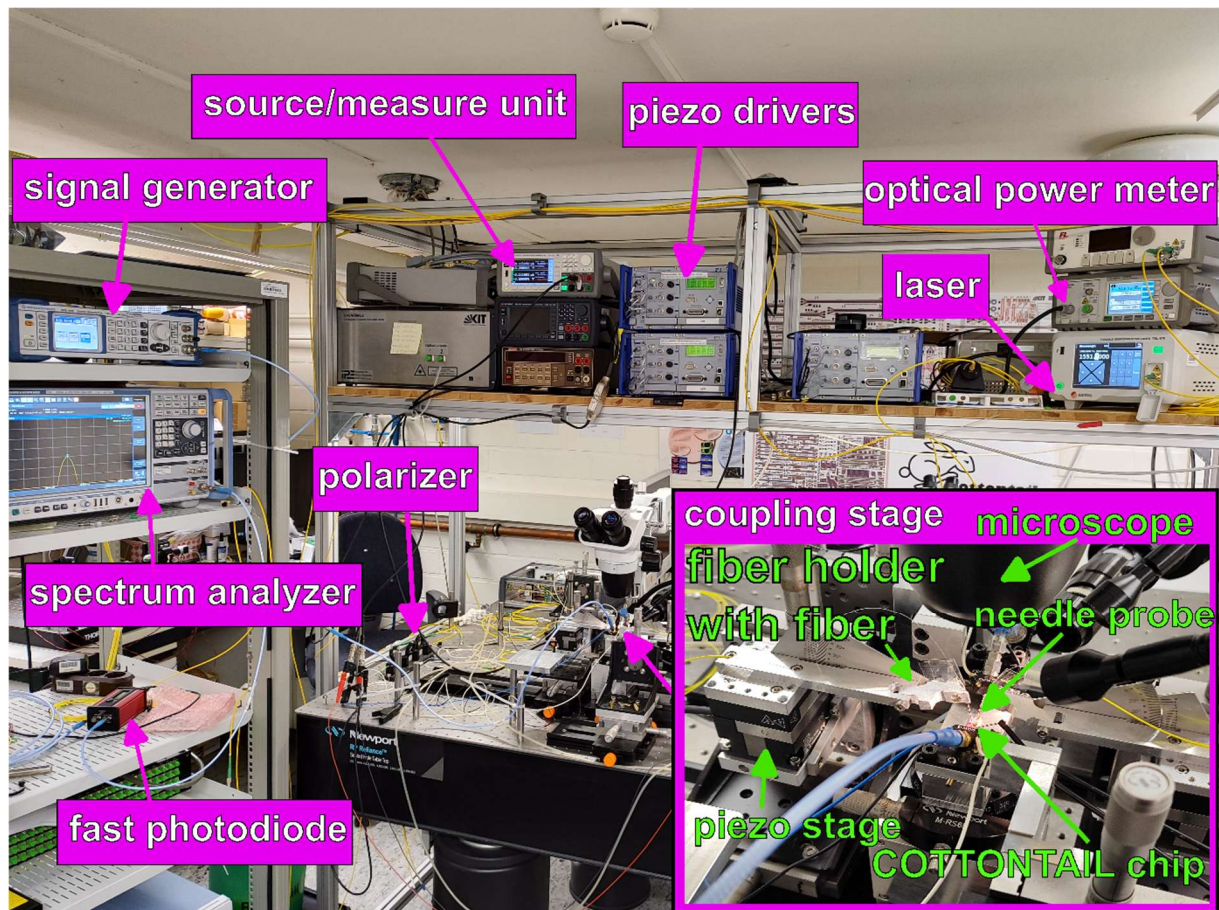
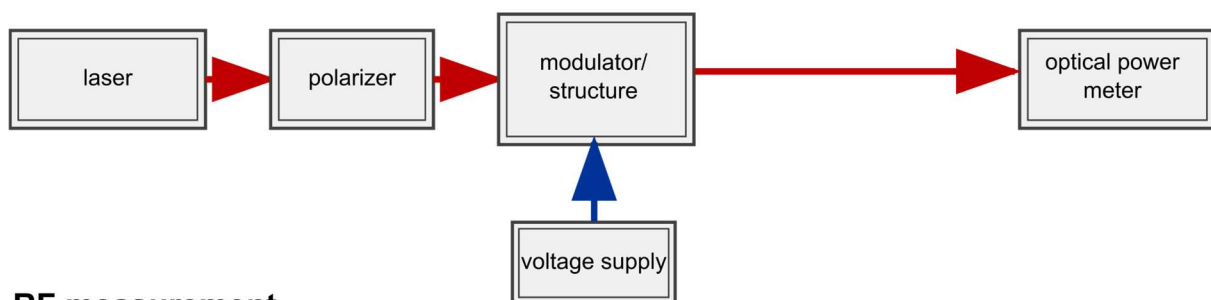


Figure 4.1: Characterization setup for DC and RF measurements with a close-up image of the coupling stage.

DC measurement



RF measurement

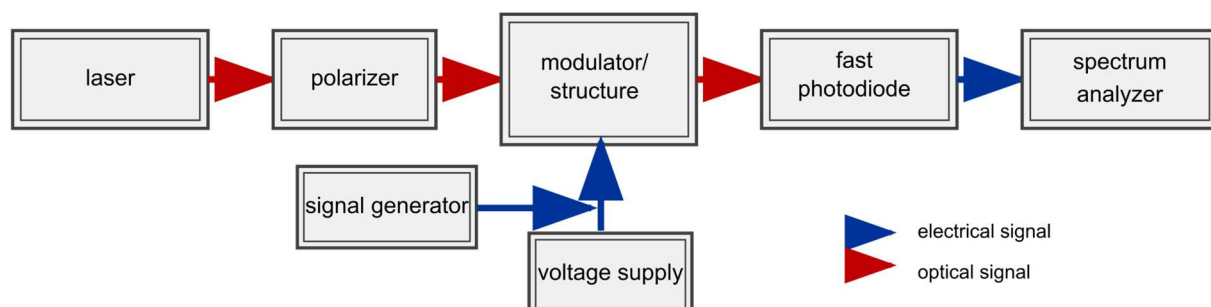


Figure 4.2: Block diagrams of the optical setup (adapted from [41]).

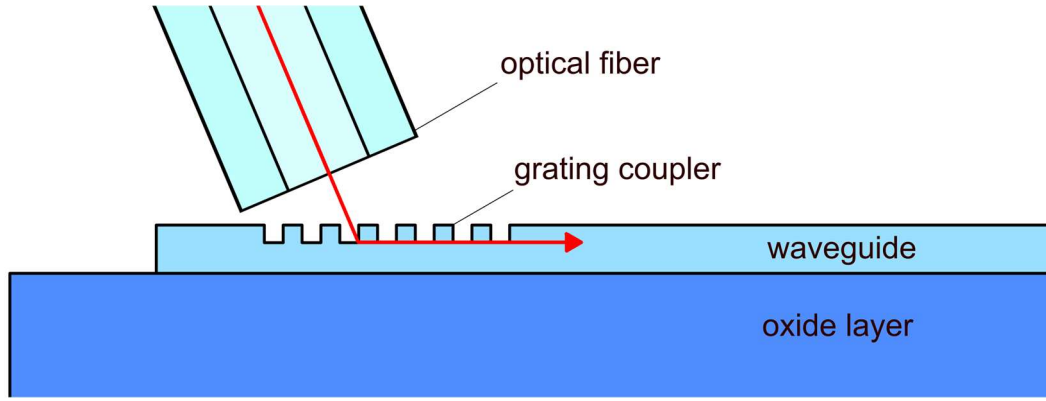


Figure 4.3: Schematic of the coupling between a flat-faceted optical fiber and a grating coupler (adapted from [12]).

If necessary, the sample/device can be connected electrically with probing needles. These are specialized RF probes e.g. the Picoprobe 40A-SG-100-EDP (GGB Industries Inc.), however in this configuration only a DC bias may be applied. If that is the case, a B2912A precision source/measure unit (SMU, Keysight Technologies) is used. The probes are positioned via micrometer screw stages above the electrical connection pads of the individual structure and then pressed onto them. It should be noted that for convenience a BT-50-400S bias-tee (Pulsar Microwave) is connected between the needle probe and the SMU (DC input in to DC+RF output out). If RM heaters are used, they are contacted in the same way and by the same devices on the opposite side of the sample.

The RF configuration replaces the final section after the coupling stage with a high-speed photodiode attached to the RF input of a spectrum analyzer. It is used to quantify the electrical-optical-electrical (EOE) response i.e. modulation efficiency. The photodiode, or more precisely the 40 GHz high-speed photoreceiver, is an RXM40AF (Thorlabs). The spectrum analyzer is an FSW43 (Rohde & Schwarz). Electrically, the B2912A from the DC configuration is kept for applying the DC-RVB. Now additionally the RF input of the bias-tee is fed by an SMB100A signal generator (Rohde & Schwarz) supplying the (sinusoidal) RF modulation signal. The latter's clock is fed by the FSW43 10 MHz reference clock to eliminate frequency deviations. All RF relevant electrical signal paths are connected via SUCOFLEX 102 high-frequency cables (Huber+Suhner) equipped with 2.92 mm connectors.

If GCs will be used in a detector framework, space constraints are very high. Due to the minimum bend radius of the fibers, standard coupling with flat fiber facets results in a high vertical profile. A fix for this is flat fiber-to-chip coupling. For the measurements presented in chapter 5 a minimal distance between the aperture of the radiation source and sample is desirable. Flat coupling relies on total internal reflection (TIR) of the guided beam at a mirror-like finished angle-polished fiber facet. Figure 4.4 (left) depicts the flat coupling. In this configuration the coupling angle is a fixed parameter. Additionally, rotational alignment is necessary with this design. So, the fiber holders with the vernier-scale shown above are replaced with fiber rotator stages (figure 4.4 - right), where the rotation angle can be set via micrometer screws. This modified setup set the previously obvious coupling angle β of the flat-faceted coupling via the polishing angle θ of the fiber. We calculate the relation between β and θ using $90^\circ = \alpha + 2\theta$ and Snell's law (2.7) [12]:

$$\theta = \frac{1}{2} \left[90^\circ - \arcsin \left(\frac{n_2}{n_1} \sin \beta \right) \right]. \quad (4.1)$$

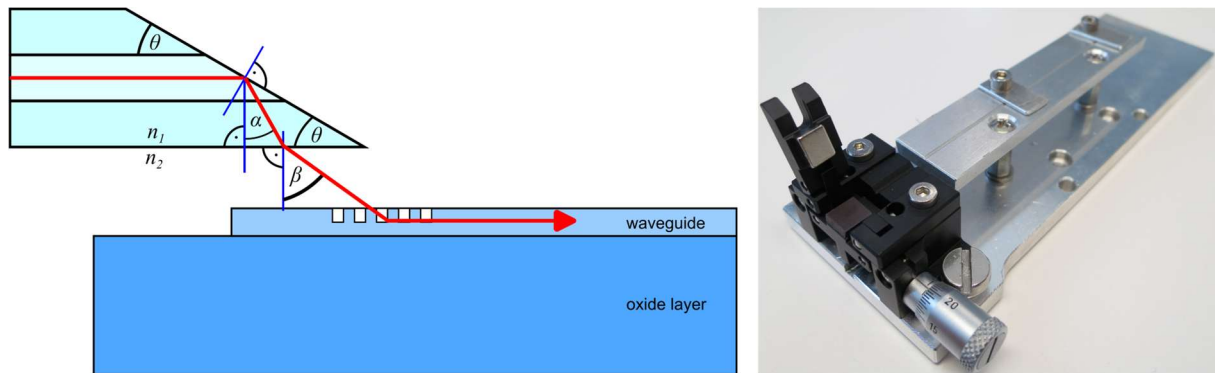


Figure 4.4: Left: Schematic of the flat fiber-to-chip coupling between an angle-polished optical fiber and a grating coupler. Right: Fiber holder/rotation stage necessary for the flat fiber-to-chip coupling in the characterization setup (both adapted from [12]).

The measurements are controlled by a Windows PC with dedicated DC and RF LabVIEW programs. The devices are addressed via GPIB and Ethernet over the SCPI protocol. In this context DC means, the DC setup configuration is used and a sweep across a wavelength interval is performed. This is also called a (DC) wavelength sweep. Among the recorded parameters the optical power received from the OPM is the main focus of this measurement. In contrast RF means, the RF configuration of the setup is utilized. But now, a fixed wavelength is set, and a predetermined interval of electrical input signal frequencies is swept over. This is called a RF frequency sweep. The diagram in figure 4.5 shows both programs' measurement routine.

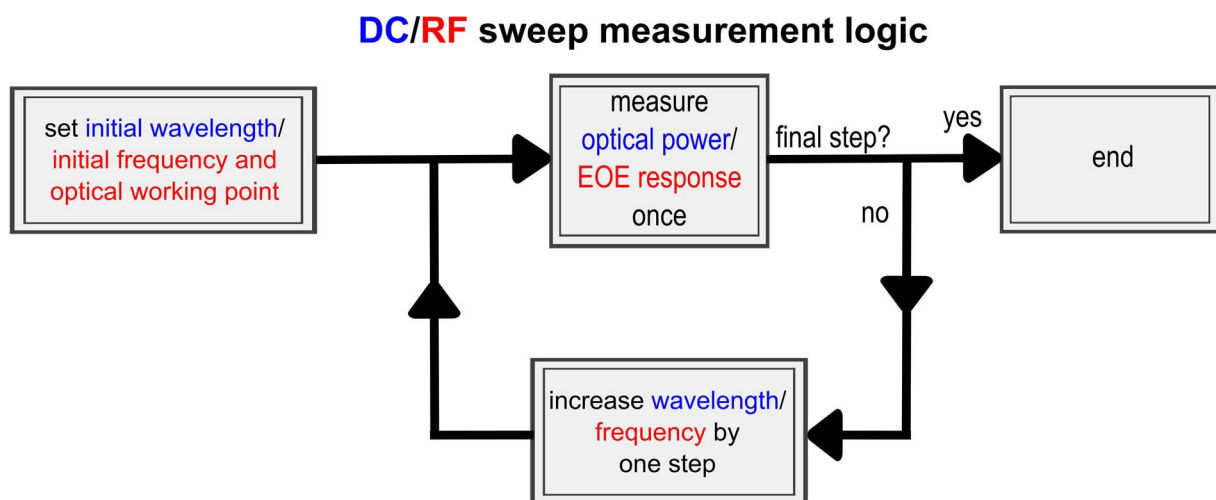


Figure 4.5: DC and RF sweep measurement logic.

4.2 Optical characteristics of the fiber-chip-coupling

For the irradiation measurements of chapter 5 we want to maximize the EOE response at wavelengths close to 1550 nm. As established in subsection 2.1.3, the coupling angle of the incident beam is crucial for efficient coupling, but it also influences the position of the maximum of the optical spectrum.

Therefore, the first optimization will revolve around the coupling angle. The manufacturer of the COTTONTAIL chip states an optimal coupling angle of 10° , but preliminary measurements have shown that the optimum is shifted to around 11° . Therefore, the goal is to find the exact optimum. This optimum also determines the polishing angle for the flat fiber-to-chip coupling.

To find the optimal coupling angle the DC configuration of the setup with the vernier-scale coupling stage utilizing flat-faceted cleaved SMF fibers is used. The measurements are done on the structure CT#D05#3_2500x0.45FCW (COTTONTAIL#wafers#chip_structure) shown on the bottom right of the top image in figure 3.2. FCW stands for focusing wideband GCs for the C-band, which are the ones used for any experiment in this thesis, unless specified otherwise. 2500 specifies the length of the waveguide between each GC in micrometers, 0.45 its width. For each measurement on both sides the fiber angle is set to a value between 7° and 15° using the vernier-scale of the fiber holders. The laser is set to -10 dBm at 1550 nm. Then the fibers are positioned above the GCs so that the optical power on at the OPM is maximized. Finally, a measurement with the DC wavelength sweep program is taken. The measurement runs between 1500 nm and 1600 nm in 0.1 nm increments.

On the left side of figure 4.6 all recorded curves are shown and on the right a selection of curves from the same dataset. Each individual curve is a set of 1001 data points. Steeper coupling angles shift a given curves peak towards higher wavelengths. The opposite is also true. Also, at lower wavelengths less noise or oscillations in the optical power are observable.

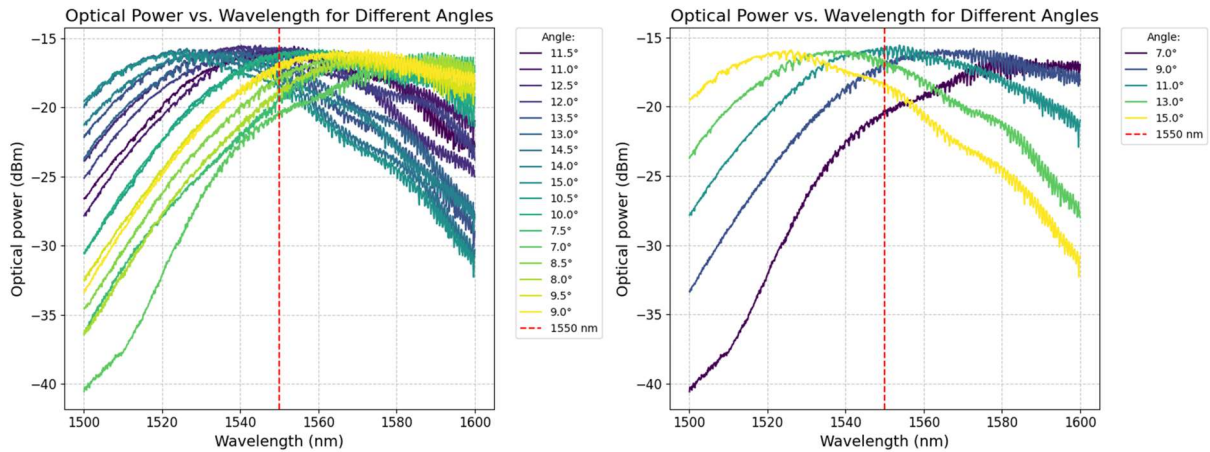


Figure 4.6: Optical spectra for different coupling angles. Left: All measurements taken. Right: Selection of representative angles.

To obtain the optimal coupling angle, the data must be processed. In the following two methods are introduced. The first searches for the position of the maximum in optical power in each curve. To combat the noise a Savitzky-Golay filter with a filter window of 151 data points and a third order polynomial is applied to each curve. From the set of resulting smooth curves each individual maximum is plotted as a blue dot in the left part of figure 4.7. Finally, a linear fit is performed on these data points, yielding the red line in the plot.

The second approach aggregates the individual measurements into a comprehensive dataset. By grouping all data points by wavelength (brown points in the right plot of figure 4.7), new distribution curves are generated, where each point represents a specific coupling angle measurement. A Gaussian

fit is applied to these distributions to determine their individual maximum positions σ . This indicates the optimal coupling angle for the most efficient coupling of the respective wavelength. As an example the resulting Gaussian fit for a wavelength of 1562.6 nm is shown on the right of figure 4.7. This procedure is applied to all 1001 wavelengths (1500 nm to 1600 nm in 0.1 nm steps), results in the green curve in the left plot.

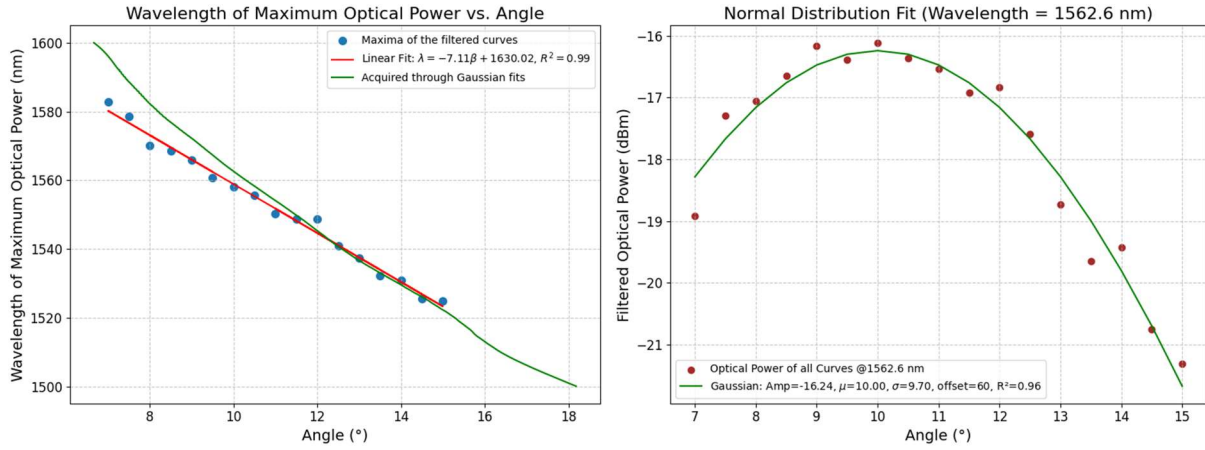


Figure 4.7: Left: Wavelength maxima of filtered curves (blue dots) with a linear fit (red line) compared to the line derived from Gaussian fits (green curve). Right: Exemplary Gaussian fit (green line) of the optical power distribution for a single wavelength 1562.6 nm across measured coupling angles (brown dots).

The linear fit as part of the first method fits the data very well with $R^2 = 0.99$. For the coupling angle range investigated, the expression

$$\beta = \frac{1630.02 \text{ nm} - \gamma}{7.11} \quad (4.2)$$

is obtained. β is the coupling angle and γ the wavelength we want to optimize the coupling for. For the target wavelength of 1550 nm we obtain $\beta = 11.25^\circ$. As for the data obtained through the Gaussian fits, it shows good agreement in the angle range from 12° to 15° . The range from 7° to 15° measurement window are also valid since it is obtained over the wavelength. But only a very limited part of a Gaussian can be fitted to the data. This explains the deviation within that range. The same explanation is true for the range from approximately 15.5° to 18.2° . This is amplified by the higher noise in the data at higher wavelengths. Due to these issues, the green curve in the left plot serves only to verify the results obtained through the first method.

Using equation (4.1) the results obtained for β can be translated to the polishing angle θ of the fiber for the flat fiber-to-chip coupling. The result of this is plotted in figure 4.8. Equation (4.2) transforms to

$$\theta \approx 0.0481 \cdot \gamma - 30.34. \quad (4.3)$$

This results in an optimal polishing angle of 41.17° for 1550 nm (using the refractive index of the Corning SMF28 fiber of $n_{\text{fiber}} = 1.462$ and $n_{\text{air}} = 1.00028$ for air).

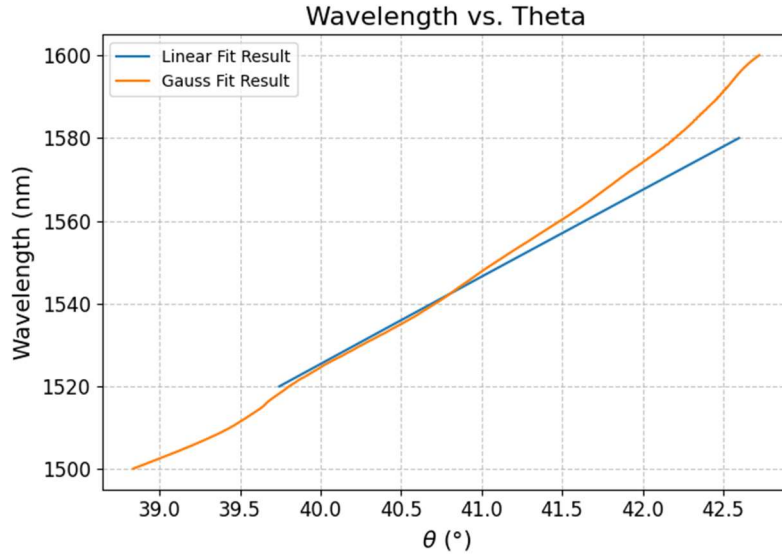


Figure 4.8: Polishing angles θ for the flat fiber-to-chip coupling and the wavelengths they produce the best coupling efficiency for. The blue line is the result of the linear fit as of the first analyzation method. The orange one results from the second method, which uses Gaussian fits on curves generated by the rearranged dataset.

The initial process for the fiber angle-polishing was developed as part of [12]. Here coupling was optimized for a different chip. The preparation process starts with gluing (glue: DELO Photobond GB310) four optical fibers of the same type into a grooved microscope slide. The slide has eight grooves sawn into it over its full length with a wafer saw. Every second groove is populated with one fiber, while the unpopulated grooves act as glue traps. This is important to trap access glue, since it creeps and the fibers are glued in consecutively. As PM fibers Corning PM 1550 polarization-maintaining panda fibers are used. If PM fibers are not needed standard SMF are glued in. Note that all of the coating and buffer layers of the fibers need to be stripped beforehand. Hence, only the bare glass fibers are polished. The gluing process is done in the setup shown in figure 4.9. For the PM fibers a rotational alignment is necessary which is achieved with a camera brought into proximity to the flat-faceted fiber already in the groove of the microscope slide. By arranging light sources around the fiber, the visible contrast between cladding glass and stress rod glass is sufficient to arrange the angle via a micrometer screw fiber rotator stage. A NI Vision Builder program helps with the exact alignment (figure 4.10). If the fiber is in its final position the glue is applied and subsequently cured by exposure to high-intensity UV light.

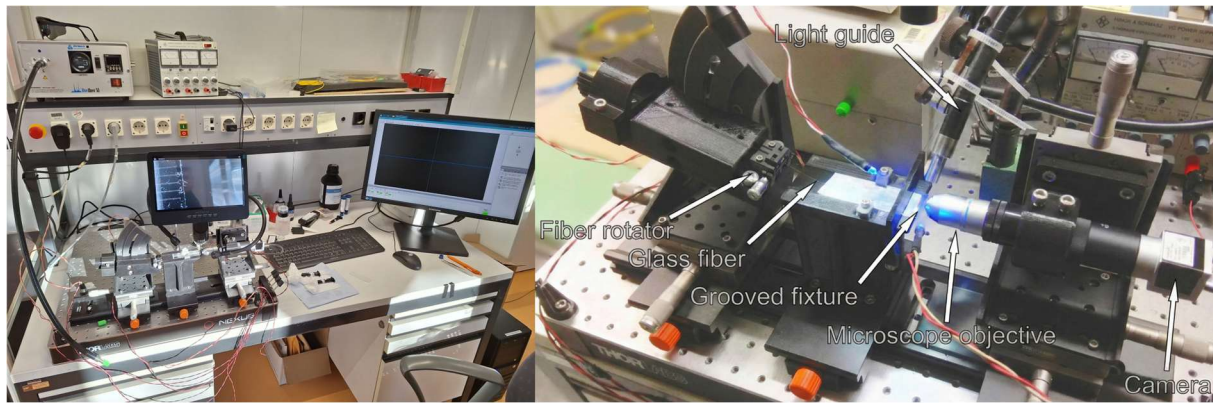


Figure 4.9: Left: The setup used for gluing the bare glass fibers into grooved microscope slides. Right: Close-up image of the setup core (adapted from [42]).

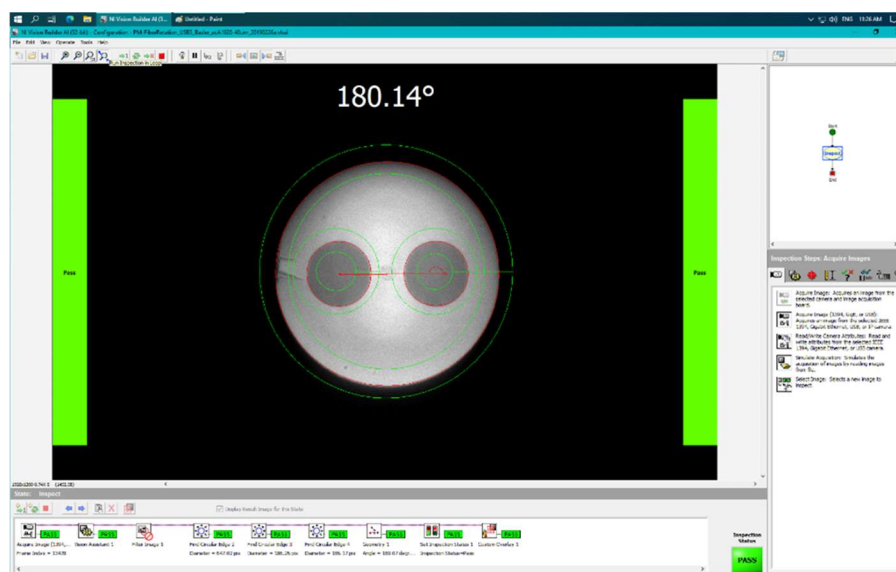


Figure 4.10: Screenshot of the NI Vision Builder application which helps with the rotational alignment. By detecting both stress rod glass cross-sections within the cladding glass, the angle can be determined. In the example shown the fiber is rotationally well aligned.

Next is the grinding step. The microscope slide with the glued fibers is mounted to a grinding chuck. This chuck is shown in figure 4.16 (left). The bottom part consists of a 3D printed carrier which has two mounting positions on a slope for a carrier stage. The slope determines the polishing angle θ . The stage that holds the microscope slide is attached to the mounting positions with a parallelogram like mechanism, which translates the chuck angle to the slide holder stage. The slide is fixed with a clamp piece tied down via screws. Additionally, a hand supporting piece on top is attached to the bottom carrier, ensuring the application of uniform pressure during the grinding process. The rough grinding and later polishing are done on a Struers LaboPol-5 grinding and polishing machine under a constant stream of water. The process starts with the rough grinding on a 1200 grid SiC sanding paper disk and ends with a specialized diamond coated polishing disks. It has a $0.1 \mu\text{m}$ grain diameter. The steps with the disks used and the grinding/polishing times are listed in table 4.1. In figure 4.11 a comparison between a successful and an unsuccessful polishing process is shown. The final control is done under a Keyence VK-X3000 3D laser scanning microscope while the fibers are still glued in. The control is limited to an optical inspection.

Step	Disk	Grinding/polishing time (with new disks)	Rotational speed
1	1200 grid SiC sanding paper	2 min	200 rpm
2	2400 grid SiC sanding paper	4 min	200 rpm
3	3.0 μm diamond lapping film	8 min	200 rpm
4	0.5 μm diamond lapping film	4 min	200 rpm
5	0.1 μm diamond lapping film	4 min	200 rpm

Table 4.1: Grinding/polishing steps with their duration and grinding disk or polishing film.

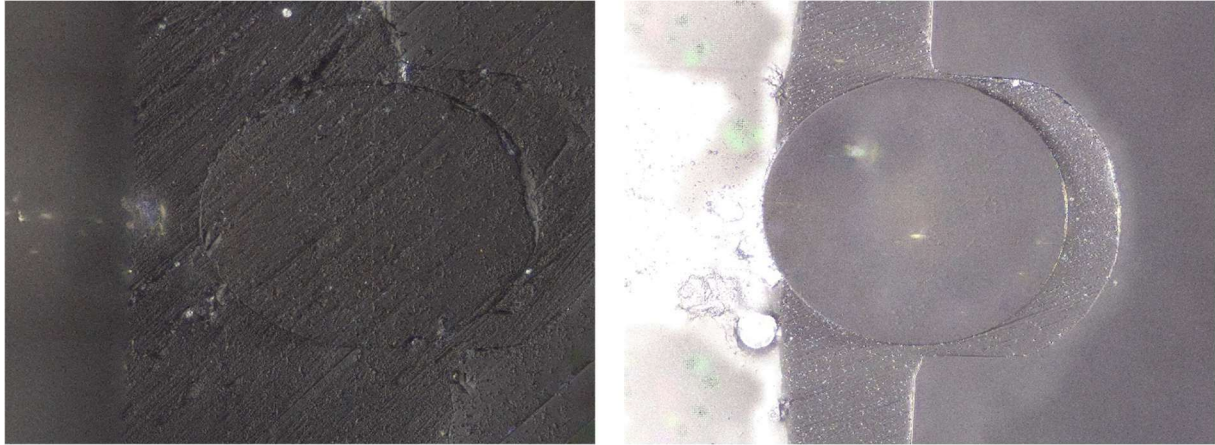


Figure 4.11: Left: Unsuccessful polishing of the fiber facet, most likely due to over-used polishing disks. Right: Well-polished fiber facet.

To make the fibers usable in the setup, they are submerged in acetone which lets the glue swell, making the fibers removeable. They are then cleaned and spliced to an appropriate fiber pigtail with an FC-APC plug on a Fitel S185PM fiber splicer. The spliced area is protected by either a heat shrink style protector or via a coating of Ultracur3D reactive urethane photopolymer (BASF Forward AM). A finished fiber-tip is shown in figure 4.12.



Figure 4.12: Side and top-side view of an angle-polished glass fiber (adapted from [42]).

The fibers are now ready to be used in the setup to check, how the chuck translates the theoretical optimal angle of $\theta = 41.17^\circ$ to the final polishing angle and if further adjustments are necessary. For that the DC configuration of the setup described in section 4.1 is used. The measurements were taken again on the CT#D05#3_2500x0.45FCW structure. In addition to the positional alignment with the micrometer screws and piezo stages and the polarization alignment, the fiber rotation in-axis must also be optimized now. To allow easier characterization of standard SMFs only the outbound side of the coupling stage towards the OPM is changed to the flat fiber-to-chip configuration. In this direction polarization does not matter anymore. The other side remains with a flat-faceted fiber at a 11.25° angle, so that the polarization for the coupling can be controlled. If a PM fiber is investigated, it will be connected directly to the laser with a PM fiber and the polarizer is skipped. In the course of the measurements with PM fibers the out-coupling fiber first was a flat-faceted one was later changed to an angle-polished one. Like before, measurements are taken with the DC LabVIEW recording program and identical settings. The laser is set to -10 dBm.

During this investigation 18 sets of four fibers were angle-polished and the majority had their transmission (DC) spectra recorded in various setup configurations. In the following a representative choice will be discussed. Table 0.1 in the Additional measurements data appendix shows all procurement variations of each set in comparison to the others.

The plot on the left in figure 4.13 depicts the transmission curve of a standard fiber polished on a 41.17° chuck. This would lead to a coupling angle of 11.23° degree, which was assumed to be the optimum. Here, the maximum of the by a Savitzky-Golay with a window length of 151 data points and a polynomial of the third order filtered curve, is at 1528.6 nm and -21.31 dBm. The measurement shown on the right of figure 4.13 has its maximum (with identical filter settings) at 1552.8 nm and -15.57 dBm. It is a PM fiber and was polished on a 42.0° chuck, which in theory should result in a coupling angle of 8.79° . The tail of the curve is flat opposed to the standard downward facing progression.

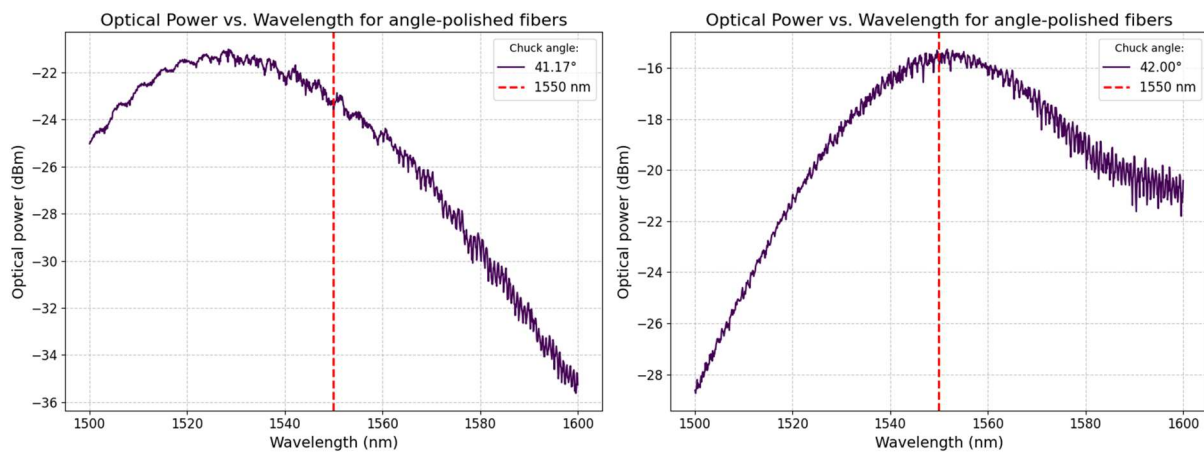


Figure 4.13: Spectrum of single fibers polished with different chuck angles. Only the out-coupling side has an angle-polished fiber, the other is flat-faceted and set to 11.25° . Left: Chuck with the theoretical best angle. Right: Chuck with an angle 42.0° that produces the best result.

The left part of figure 4.14 shows the DC curve of a setup with the inbound and outbound fibers being angle-polished. The inbound PM fiber was angle-polished on a 41.17° chuck (11.23° theoretical

coupling angle) and the outbound standard SMF on a 42.0° one (8.79° theoretical coupling angle). The (filtered) maximum of this constellation is at 1552.7 nm and -17.84 dBm. The right side of figure 4.14 shows the transmission curves of five PM fibers, all angle-polished with the identical 42.0° chuck.

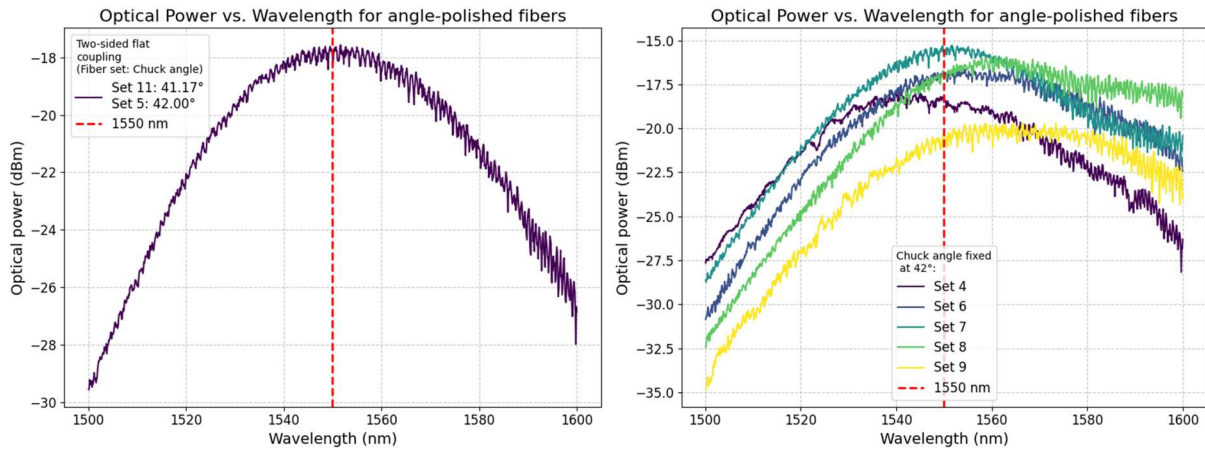


Figure 4.14: Results of single-sided angle-polished fiber measurements. Left: Best both-sided flat fiber-to-chip coupling result. Right: Variance between different sets grinded with the same chuck (angle). Variance due to chuck wear and non-uniform pressure application during the polishing process.

The measurements show that the chuck angle doesn't directly translate to the polishing angle θ of the fibers and even for identical parameters the results show variance. Generally, the polishing angle turns out smaller, therefore leading to a bigger value of the coupling angle β . The plot on the left in figure 4.13 shows this clearly. To compensate for this divergence the 42.00° chuck for early sets delivers the best result. This means that an additional polishing angle of 0.83° is applied. This addition translates over proportionally to the coupling angle by a decrease of 2.44° . As the right plot in figure 4.14 shows, angle-polishing the fiber facets reliably to a target angle is a complex undertaking. An identical setup may lead to perfect, low or high angles. Next to the chuck angle there are several strong influences; wear and play of all movable 3D printed plastic parts, which is stronger than expected, wear on the grinding and polishing disks and the force and its uniformity applied during the grinding and polishing process. Also, the optical inspection itself is not enough to ensure a high surface uniformity and quality of the facet. Even if the facet appears to be well-polished, their spectra still may be shifted and warped like shown on the left of figure 4.15. The in-set variance of the fibers is relatively low as can be seen on the right of figure 4.15.

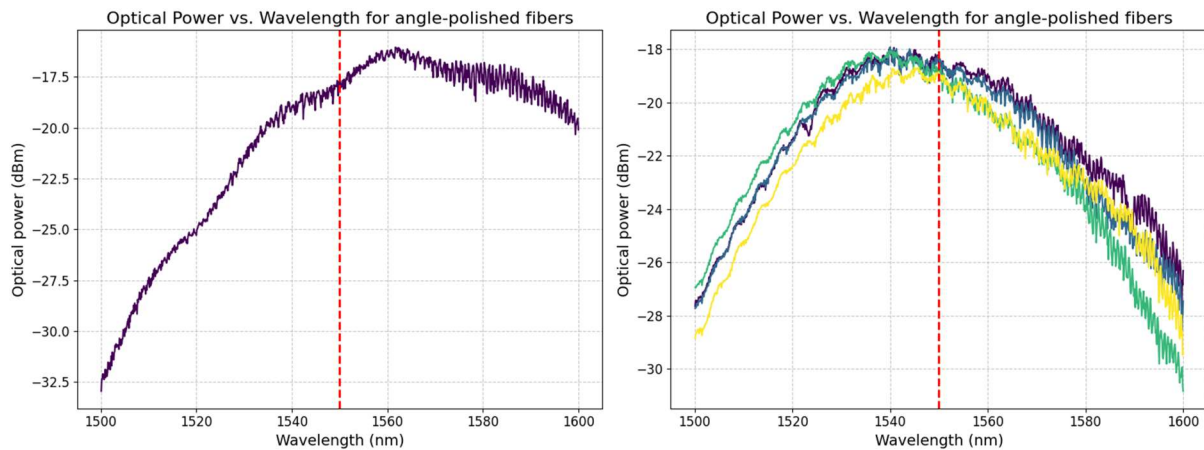


Figure 4.15: Left: Spectrum of an angle-polished fiber that passed visual control but nonetheless produces a warped spectrum. Right: Spectra of all four angle-polished fibers within one set.

To combat these issues, a more rigid and wear resistant grinding chuck construction was developed. As a first step the chuck remained the same, however the parallelogram like microscope-slide-stage was constructed from metal. The combination of plastic 3D printed with machined metal parts proved to be error prone. Therefore, a new chuck was conceived. Now it only serves as a stop angle, where the parallelogram construction, completely made from aluminum, is anchored to. The chuck setup now is more wear resistant and less prone for bending. Leaving the microscope slide, which carries the fibers, as the suspected dominant source for uneven facets, if uniform force is applied during the grinding and polishing process. The three chuck iterations are shown next to each other in figure 4.16. A new series of angle optimization measurements with this new chuck are warranted. However, due to time constraints these measurements are not subject of this thesis.

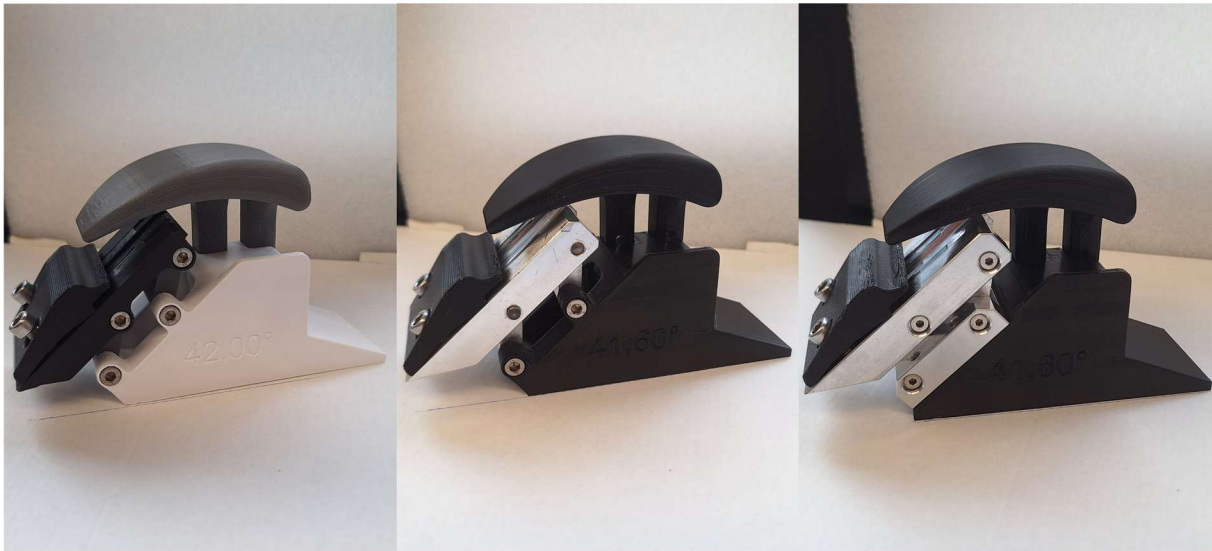


Figure 4.16: Chucks for angle-grinding/polishing glass fibers glued into a grooved microscope slide. Left: Full 3D printed plastic version. Middle: The stage and arms of the parallelogram mechanism have been replaced with aluminum parts. Right: The full mechanism is made of aluminum. The 3D printed main part only serves as a stop to determine the polishing angle.

4.3 Optical characteristics of the COTTONTAIL chip

On top of the fiber-to-chip coupling, also some optical aspects of the chip itself have been investigated. The first two can be analyzed in the same measurement campaign, the waveguide and GC losses. Next to the general interest in these quantities, they are paramount to judge the coupling efficiency of permanent sample builds for any measurement.

The measurements are conducted with the DC configuration of the setup with the coupling stage for flat-faceted fibers. The DC LabVIEW program takes measurements with the same settings as in section 4.2. Since the measurements were conducted before the coupling angle optimizations the previously suspected best coupling angle of 11° was used. Again, the coupling was optimized with the micrometer screws and piezo stages by maximizing the transmitted optical power. Table 4.2 and figure 3.3 show all the waveguides which were investigated. The measurements were repeated twice for each structure. The absolute dopings of the waveguides are not known. Relative to the background doping concentration of the standard waveguide, there are p^+ - high positive, n^+ - high negative, p^{++} - very high positive and n^{++} - very high negative. The laser was set to 0 dBm.

Structure name	Waveguide length [μm]	Relative doping	GC	Chip	Laser settings
TSNoneSpiral1	2000	None	FCW	CT#D05#3	1550 nm at 0,dBm
TSNoneSpiral3	4000	None			
TSNoneSpiral4	6000	None			
TSNoneSpiral2	8000	None			
TSNoneSpiral5	10000	None			
TSNBODYSpiral2	1000	n^+			
TSNBODYSpiral4	2000	n^+			
TSNBODYSpiral3	3000	n^+			
TSNBODYSpiral1	4000	n^+			
TSPBODYSpiral2	1000	p^+			
TSPBODYSpiral4	2000	p^+			
TSPBODYSpiral3	3000	p^+			
TSPBODYSpiral1	4000	p^+			
TSNPLUSSpiral1	500	n^{++}			
TSNPLUSSpiral2	1000	n^{++}			
TSNPLUSSpiral3	1500	n^{++}			
TSPPLUSSpiral3	500	p^{++}			
TSPPLUSSpiral2	1000	p^{++}			
TSPPLUSSpiral1	1500	p^{++}			

Table 4.2: Waveguides on the COTTONTAIL chip used for the waveguide and grating coupler loss analysis.

The results of the measurements are shown in figure 4.17 and in the appendix Additional measurement data section in figure 7.2. They show all waveguide length data points for a given doping level. For both measurement runs a linear fit was made. The slope of the straight is equal to the loss of optical power per length unit (here $\text{dBm} \cdot \mu\text{m}^{-1}$) of a given waveguide. The intersection of the y axis corresponds to the

loss of both GCs. For example, according to the data plotted figure 4.17. The first run yields a standard waveguide attenuation of $1.5 \cdot 10^{-4} \text{ dB} \cdot \mu\text{m}^{-1}$ and a GC attenuation of 3.09 dB. The second run results in $1.7 \cdot 10^{-4} \text{ dB} \cdot \mu\text{m}^{-1}$ and 2.82 dB respectively. All results are included in their respective figures.

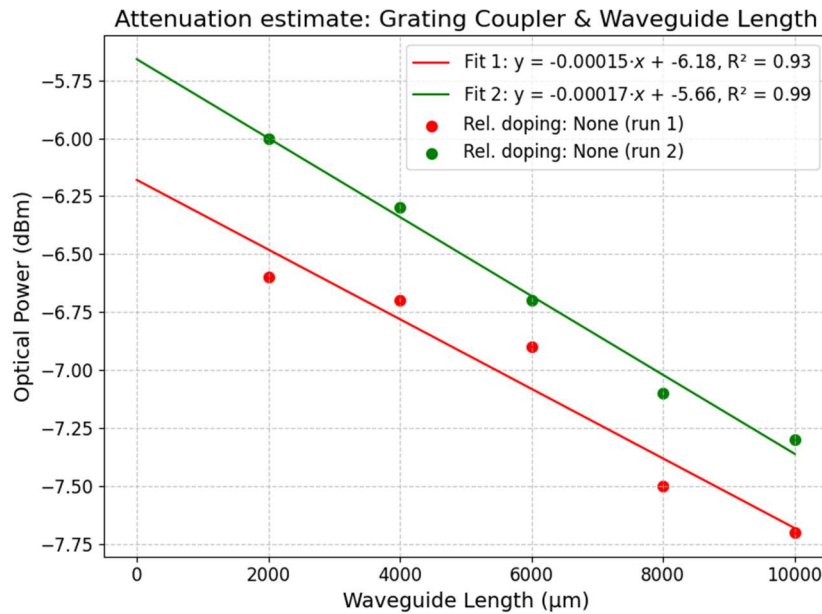


Figure 4.17: Analysis of the waveguide and grating coupler losses of the in relative terms undoped waveguides as of Table 4.2.

Regarding the standard waveguides shown in figure 4.17, both runs align reasonably well. Since run two delivers a better fit quality and lower GC losses, it is considered the most reliable estimate. Their slight disagreement can be possibly explained by a maladjustment in the polarization alignment via the polarizer, worse positional alignment due to drift sometimes effecting the OPM and piezo stages and contamination on the fiber facets at the coupling stage. Concerning the measurements of the doping variations in figure 7.2, they do not provide a concise image. Since identical GCs were used the fits should intersect the y-axis much closer to the results of the standard waveguides. Since this is not the case, it must be suspected that the MFD of the waveguides and GCs are mismatched or the doping across the structures of the same doping type are not constant. Another possibility is that the manufacturing process does not allow for a prompt transition between a GC made for an undoped waveguide to a modified one with very high doping concentration. The second “n+” run shown in figure 7.2 with the very high negative doping concentration, delivers a good fit quality, which may hint into that direction combined with the MFD mismatch. However, due to the low sample size this could be a coincidence.

With these estimations for waveguide and GC losses, we can predict a best-case coupling scenario for the RMs shown in figure 3.4. This is of interest to judge the sample builds shown later in subsection 4.5.2. Their bus waveguides are roughly 500 μm long and they use the standard FCW GCs. Therefore at 1550 nm and a laser output power of 0 dBm the optimal suspected power receivable at the OPM is −5.75 dBm. It needs to be considered that waveguide spirals have been measured here. The mode field in the curves of the spirals is pushed further into the cladding, where losses most likely increase compared to straight or only slightly bent waveguides. As a reference, the filtered data of the measurements in the previous section 4.2 delivers a value of −15.95 dBm at 11° coupling angle at 11.5° −16.12 dBm at 1550 nm. It can be expected that at the optimum of 11.25° the value would be

slightly higher. Since there the laser was set to -10 dBm both values can be increased by 10 dBm. This yields -5.95 dB and -6.12 dB respectively as the minimum possible attenuation, reinforcing the estimate for the best-case coupling of -5.75 dB.

In the following a final aspect of the COTTONTAIL chip is analyzed: Its tunability. For future use of the SiPh technology, working point control i.e. tunability is paramount. Especially RMs, which are the technology most likely to be implemented in detector data transmission systems, are sensitive to changes in temperature. An increase in temperature increases the effective refractive index of the resonator, changing its optical length and shifts the notch positions in its spectrum. Thermal expansion slightly increases the shift. This can be used via the heater to control notch positions and therefore the working point position, which may lie on any slope of a given notch. A control could stabilize the working point at a fixed wavelength. For a temperature range of -20 °C to 80 °C the behavior can be approximated to be linear [19].

Measurements are taken in the DC configuration of the setup with the flat-faceted fiber-to-chip coupling stage. The needle probes are connected to the heater contact pads to the right of the CT#D05#RingMod1 (COTTONTAIL#wafers#chip_structure) RM shown on the right in figure 3.4. No DC bias to the pn-junction of the RM was applied. The laser was set to -10 dBm, limiting the heat input through optical losses. For each step, a DC sweep was conducted while the voltage supplied to the heater was increased by 0.1 V relative to the previous measurement. The heater voltage spanned a range from 0 V to 3 V. A total of 31 measurements were recorded. To achieve this the LabVIEW DC program offers an automated function. Every wavelength data point has an accompanying voltage and current reading from the SMU. The heating power applied can be calculated by multiplying both.

Figure 4.18 shows a selection of all recorded curves and figure 7.3 in the Additional measurement data section shows all. The color gradient from blue to red signifies an increase in heating power. By following a given notch along the gradient iterations, a shift larger than one FSR can be observed. With increasing heating power the overall optical power of the curves decreases.

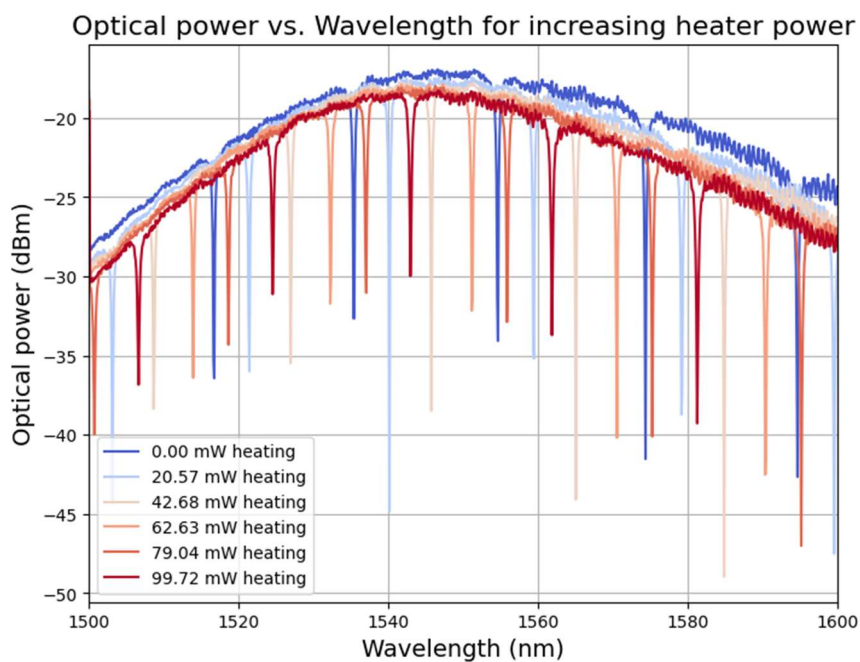


Figure 4.18: A selection of transmission spectra of a heated ring modulator.

The tracking of the notches of the measurement with no applied heating power over the heated range results in figure 4.19. Also, an example for the heating power needed to shift a notch by one full FSR is shown. Note that the FSR, as mentioned in 2.1.4, is not expected to be a constant value. However, excluding notches one and seven (as of the plot), formulating the average heating power for one FSR shift might be useful. Excluding the FSR values between notches one and two and six and seven the average amounts to 72.23 mW. The reason for excluding the FSR values containing notches one and seven is that for their fits too few data points are available introducing uncertainty. The FSR here is calculated exclusively using the fitted lines. The FSR values chosen to calculate the average are the ones at 0 mW heating power. A heating slope can be formulated, again excluding notches one and seven, which yields

$$\lambda \approx 0.267 \text{ nm} \cdot \text{mW}^{-1} \cdot P + \lambda_{\delta T=0}. \quad (4.4)$$

P is the heating power and $\lambda_{\delta T=0}$ the initial wavelength of the notch. The decreasing optical power measured at increasing temperatures as seen in figure 4.18, can be explained by higher losses via phonons or a decreasing coupling efficiency. The latter is plausible, as the piezo stages exhibit a drift over time.

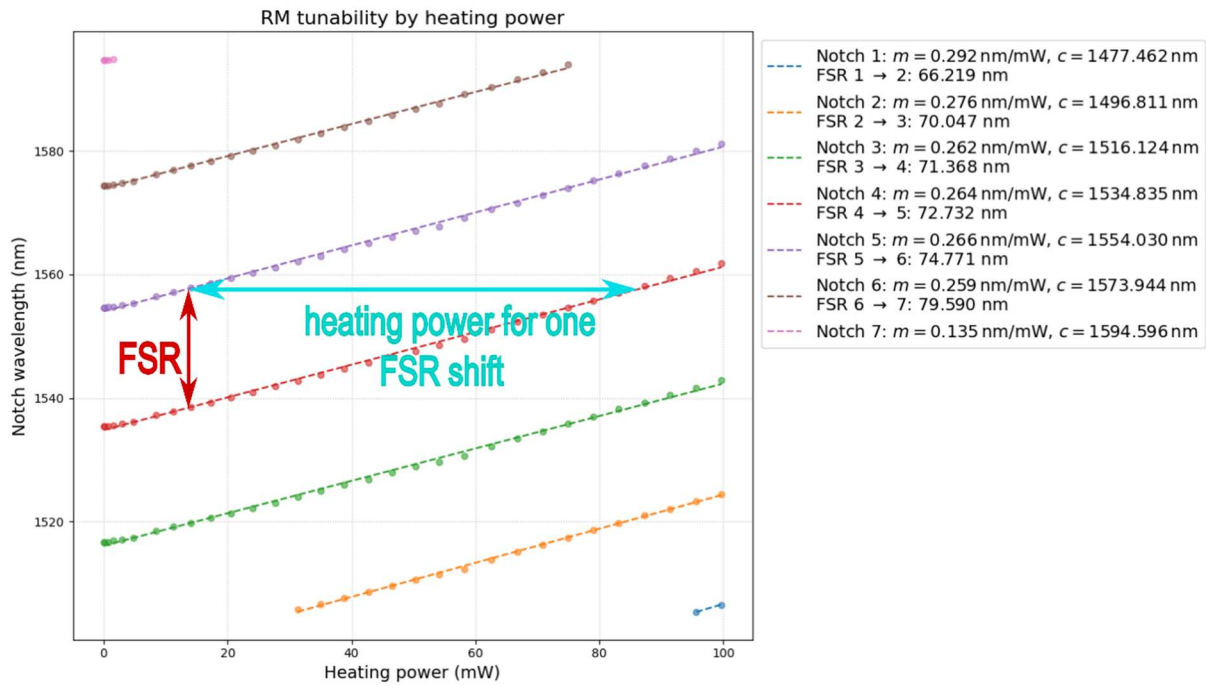


Figure 4.19: Tunability of a COTTONTAIL ring modulator: The vertical distance at a given heating power between two neighboring lines is the FSR. For the given FSR example (red arrow) the blue arrow shows the necessary heating power to tune the ring modulator by the FSR distance.

4.4 Fan-out PCBs

To enable quasi-permanent electrical connections to the structures on the COTTONTAIL chip for the in-situ measurements discussed in chapter 5 so-called fan-out PCBs are necessary. They act as a bridge between a physical connector, to which cables can be attached to, and the pads on the chip. The pads

are wire bonded to the fan-out PCBs. Opposed to DC or low frequency electronics, RF electronics come with much higher requirements to the materials, components and geometry as shown in section 2.3.

As a part of [35] microstrip and grounded coplanar waveguide (GCPW) PCBs, manufactured by Multi-Circuit Boards Ltd., using FR4 and Rogers 4350B base material, were compared. Typically, due to their advantageous ϵ_r the latter is preferred for RF applications. However, it was found that FR4 is a viable option for short straight waveguides. Due to the lower price, faster delivery time, simplicity of the PCB designs and low stack height (1.0 mm) the FR4 based four layers 4L-02 process was chosen (figure 4.20) [43]. The stack height is important, since if it is comparable to the COTTONTAIL chip height, it shortens the wire bond lengths. This is advantageous since impedance matching between them is usually not feasible, apart from adjusting the rough distance between single wire bonds. Therefore, limiting their length is common practice. To reliably wire bond to the PCBs the copper surface finish (electroplating) is set to be bonding gold (ENEPIG).

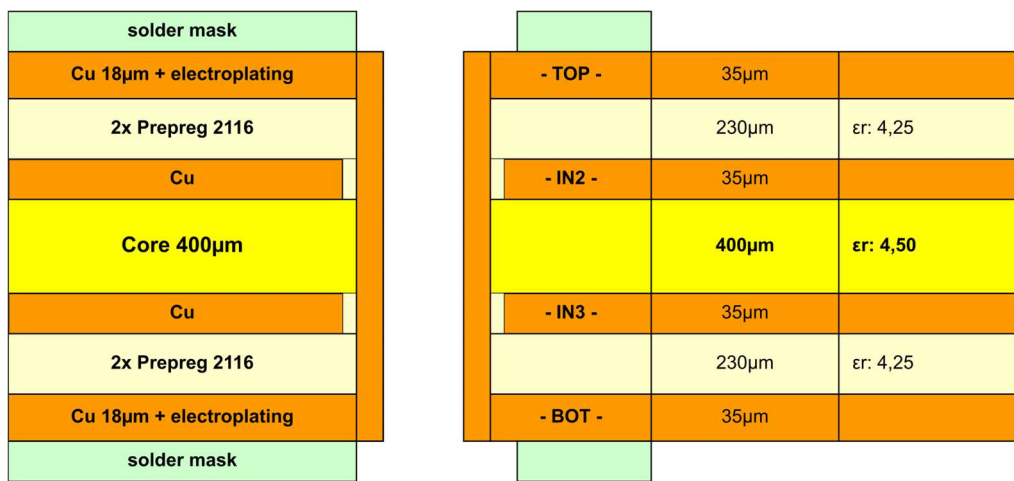


Figure 4.20: Layer diagram of the four layer 4L-02 process by Multi-Circuit Boards (adapted from [43]).

The manufacturing process determines most of the boundary conditions for waveguide geometries. As microstrips are not ideal, because proper grounding is required for both the MZMs and the RMs, it was chosen to use coplanar waveguides (CPWs) and GCPWs. To achieve an impedance matched geometry for both waveguide types, suggested geometries from TX Line 2003 (Applied Wave Research Inc.) and the KiCad EDA were compared to the best results of [35]. Their results are shown in table 4.3. All non-predetermined parameters are the waveguide width W and the gap width between waveguide and ground plane on the same layer G . Note that for the CPW design the inner two layers are not populated by copper, but the opposed outer layer still is. For simplicity, it was assumed that the resulting gap of 930 µm is enough not to influence conduction properties noticeably. Also, the FR4 core has a dielectric constant of 4.5 and the Prepreg 2116 only 4.25. Since most of the electric field should be confined to the Prepreg, its value was chosen. Realistically the value should be slightly larger. For the GCPW these considerations are irrelevant since the two inner layers act as a ground confining the field only within the Prepreg layers. The length of the conductor also plays a role in this process. At the point in time when the PCB designs were started the best guess was 1.7 cm. This number comes from half of the width of a standard microscope slide, which are used for the sample builds shown in section 4.5 and the electrical connector depth.

Waveguide type	TX Line 2003	KiCad	Previous best [35]
CPW	$W = 680 \mu\text{m}$ $G = 110 \mu\text{m}$	$W = 267 \mu\text{m}$ $G = 110 \mu\text{m}$	None
GCPW	$W = 290 \mu\text{m}$ $G = 110 \mu\text{m}$	$W = 295 \mu\text{m}$ $G = 110 \mu\text{m}$	$W = 315 \mu\text{m}$ $G = 130 \mu\text{m}$

Table 4.3: Suggested coplanar and grounded coplanar waveguide geometries to achieve impedance matching to 50Ω .

It was decided to follow TX Line 2003's suggestion for the CPW designs. For the GCPW designs it was chosen to follow the experiences of [35].

The electrical connection to the measurement equipment for all designs is realized via SV Microwave 1521-60201 solderless compression mount 2.92 mm RF connectors. The connectors were chosen since they can be reused and they were already investigated in [35]. There it was found that the manufacturer's recommended footprint is not the best possible option. A straight continuation of the waveguide center geometry to the edge of the PCB delivers better results.

Before specific designs are introduced, the taper footprint towards the wire bonds, as commonality to all fan-out boards should be highlighted. Again, the standard 4L-02 process sets design constraints. To achieve the best possible transition between PCBs and wire bonds the geometries should adiabatically transition into each other. The smallest possible footprint apart from the layer height is a waveguide of $W = G = 100 \mu\text{m}$. The MZM pads, on the other hand, have a width of $60 \mu\text{m}$ and a pitch of $100 \mu\text{m}$. Therefore, to keep wire bond length as short as possible, the smallest allowed geometry has been chosen. This refers to the layout around the inner waveguide. The ground planes around are designed to be at least five times wider than the inner layer at any time [44]. This is a common rule of thumb for CPWs and GCPWs. But the planes may end up wider as the highest priority is to avoid corners close to the inner waveguide. The resulting tapers can be seen in figure 4.21. It should be noted that the tapers themselves are not impedance matched. This decision was taken to speed up the design process. The tapers on the COTTONTAIL chip from the bond/connection pads towards the individual structures are also not impedance matched, due to chip design restrictions.

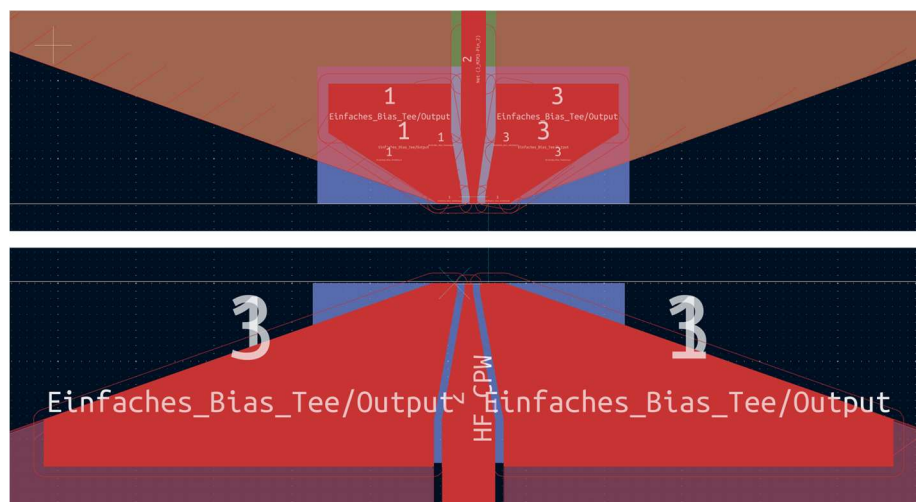


Figure 4.21: Top: Grounded coplanar waveguide taper layout. Bottom: Coplanar waveguide taper layout. Both are in scale with each other.

The full layout is shown in figure 4.22. The top part shows the front side with the top copper/ENEPIG layer. The lower part is the bottom copper/ENEPIG layer. Although the PCB layout is symmetrical on both sides, the solder mask design differs in the capacitor region for the MZM designs. The layouts were designed with standard footprints for the capacitors. Therefore, their pads are not fully surrounded (and connected) by Cu and two footprints have a certain distance between them. Given the advantages of high-density capacitor placement in RF applications, a modified layout was designed for each MZM. Standard footprints were replaced by a continuous solder mask opening over the capacitor region. This exposes the copper planes, separated only by a minimal gap, to facilitate low-impedance and dense capacitor mounting. These modified layouts were placed on both sides of the PCB. A view of the complete PCB is provided in figure 4.23. There the top row of layouts only contains designs with dedicated capacitor footprints (except for three RM layouts to the right). The middle row mostly consists of modified designs.

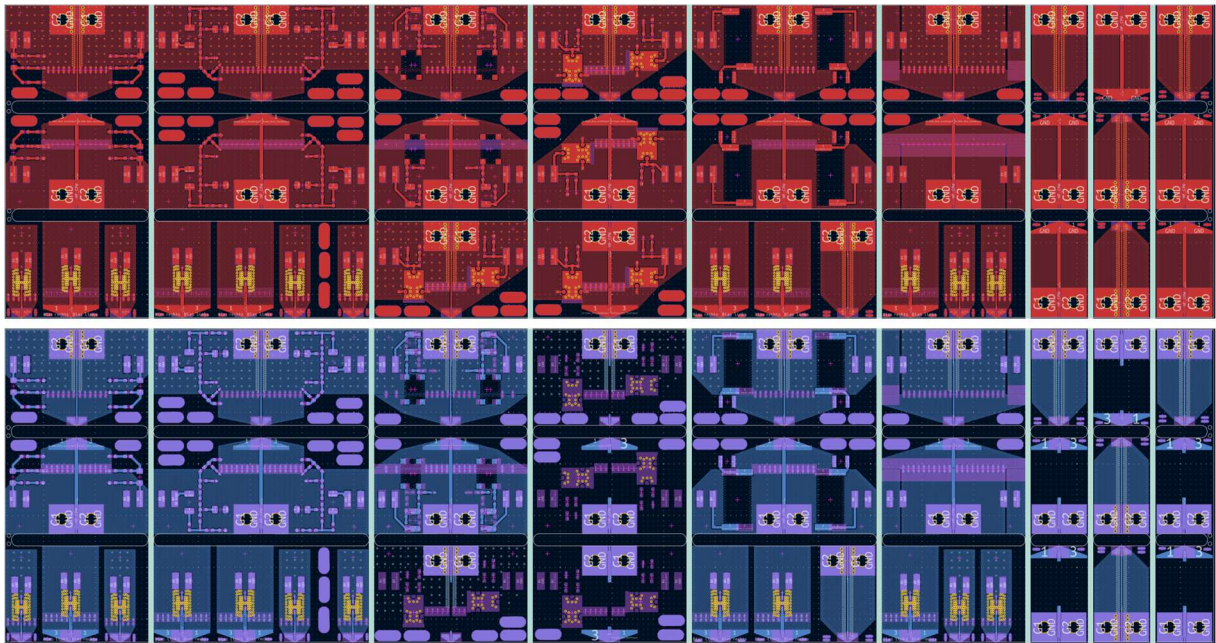


Figure 4.22: Top (red) and bottom (blue) Cu layer of the full PCB with all fan-out designs.

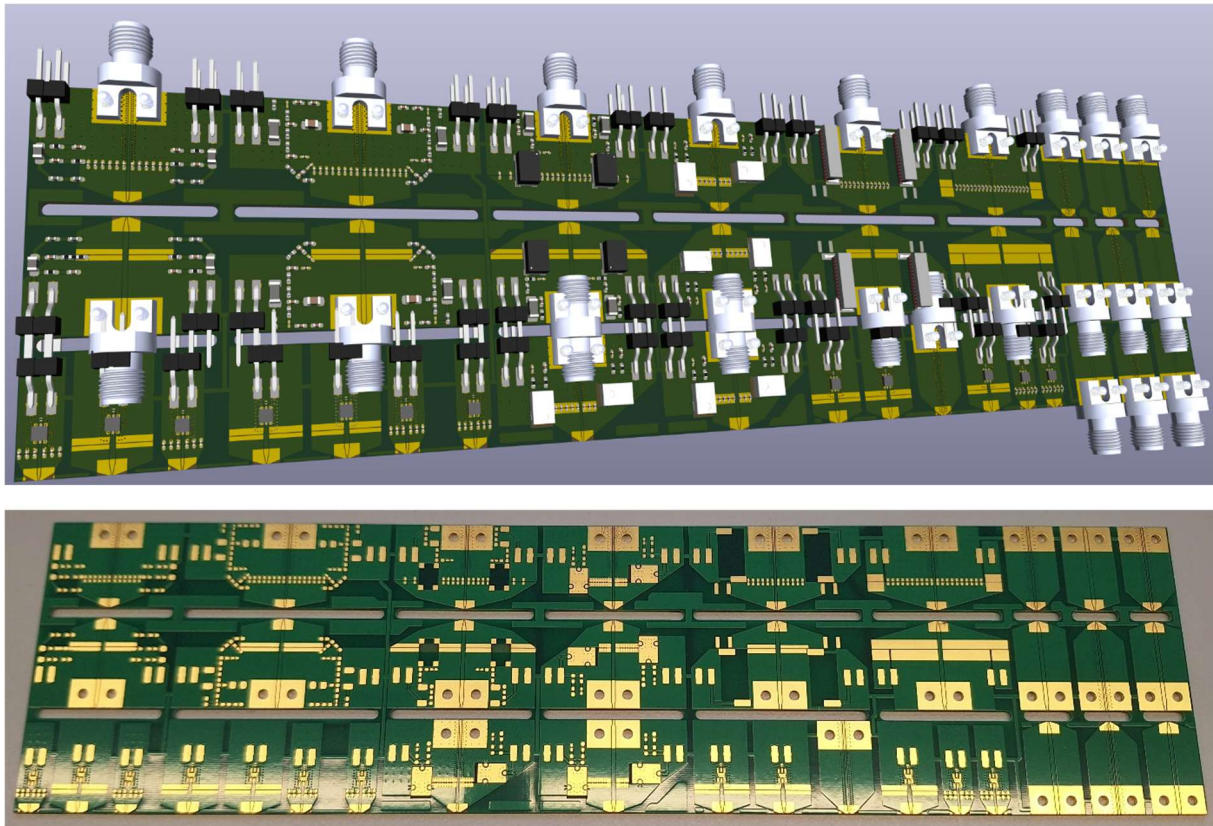


Figure 4.23: Top: Simulated top view of the full PCB. Bottom: Real image of the top side of the full PCB.

4.4.1 Fan-out PCBs for ring modulators

The final layout of the fan-out PCBs for the RMs are shown in figure 4.24. No electrical components apart from the RF connector ought to be mounted. The GCPW design has, next to the screw holes for the RF connector, two rows of through hole vias (200 μm hole diameter and 400 μm pad diameter). They are positioned as close together as allowed by the design rules i.e. a pitch of 400 μm . This practice is called via fencing and serves to reduce the loop inductance of the returning signal, thereby improving impedance stability. Additionally, the vias interconnect all ground planes, preventing unwanted substrate modes propagating through the dielectric layers. The planes around the inner RF waveguide are grounded through the connector. The RMs themselves only have two connection pads. So, only one of the ground planes will be connected to a given RM. It must be assumed that this adds a penalty to the RF transmission.

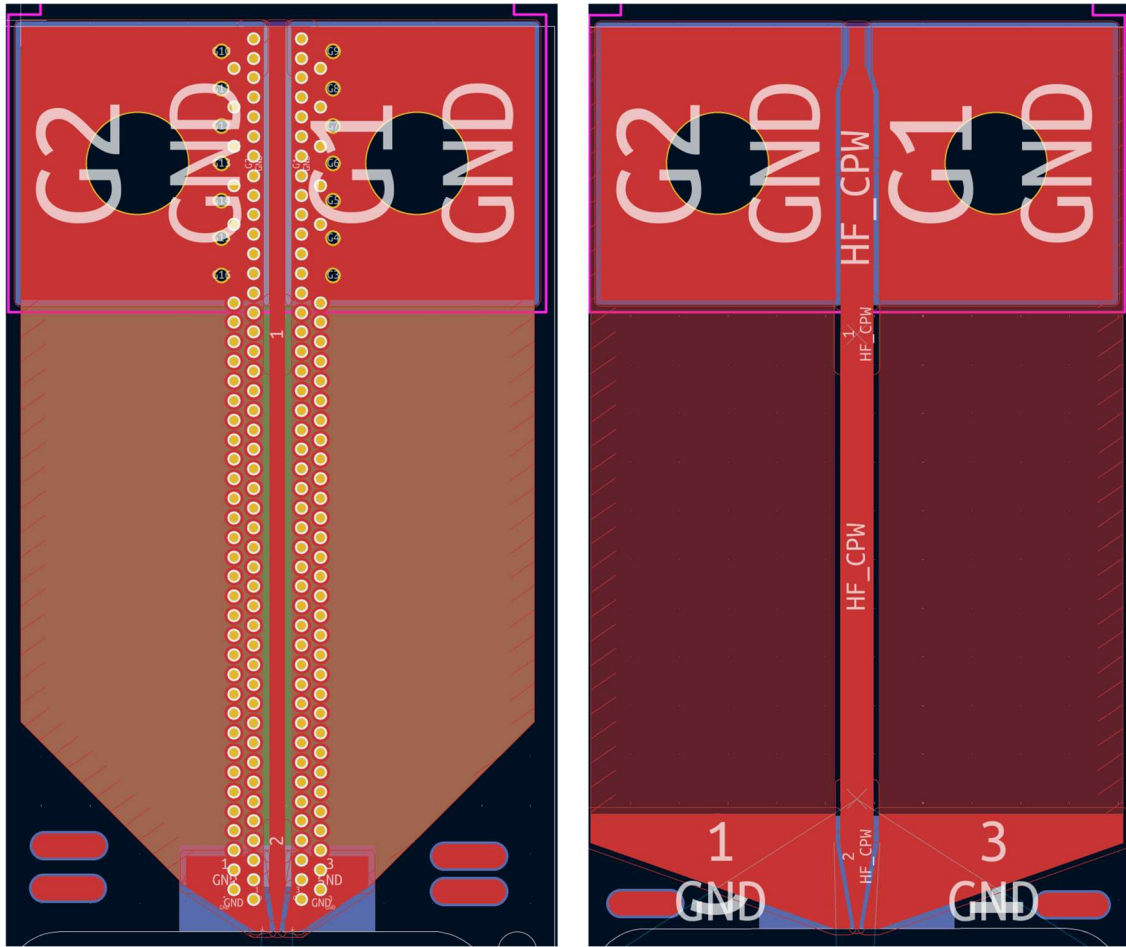


Figure 4.24: Left: Grounded coplanar waveguide fan-out layout for ring modulators. Right: Coplanar waveguide fan-out layout for ring modulators.

To investigate the possible penalty and to evaluate the design in general an experiment with the finished boards was conducted, where the S -parameters could be analyzed. For that the fan-out PCBs were singulated with a pneumatic PCB depaneller and a PCB guillotine. Then two of the same type were glued facing each other on a microscope slide using OP-54 optical adhesive (Dymax). The glue was cured under intense UV irradiation. Next the PCBs were wire bonded (wedge bond shapes) to each other. For each waveguide type a configuration with two and three wire bonds was created. Finally, RF connectors were screwed on all ends. The finished samples are shown on the left in figure 4.25.

With these preparations the conductors can be connected to the measurement setup (right side of figure 4.25). To measure the S -parameters each configuration is connected to a Rohde & Schwarz ZNA 43 vector network analyzer from both sides with VNAX-2FT-KMVRF+ cables (Mini-Circuits). Beforehand the setup was calibrated with a $50\ \Omega$ impedance calibration kit. Measurements were taken from 10 MHz to 43.5 GHz in 5 MHz steps, leading to 8699 data points. The intermediate frequency bandwidth was set to 100 kHz.

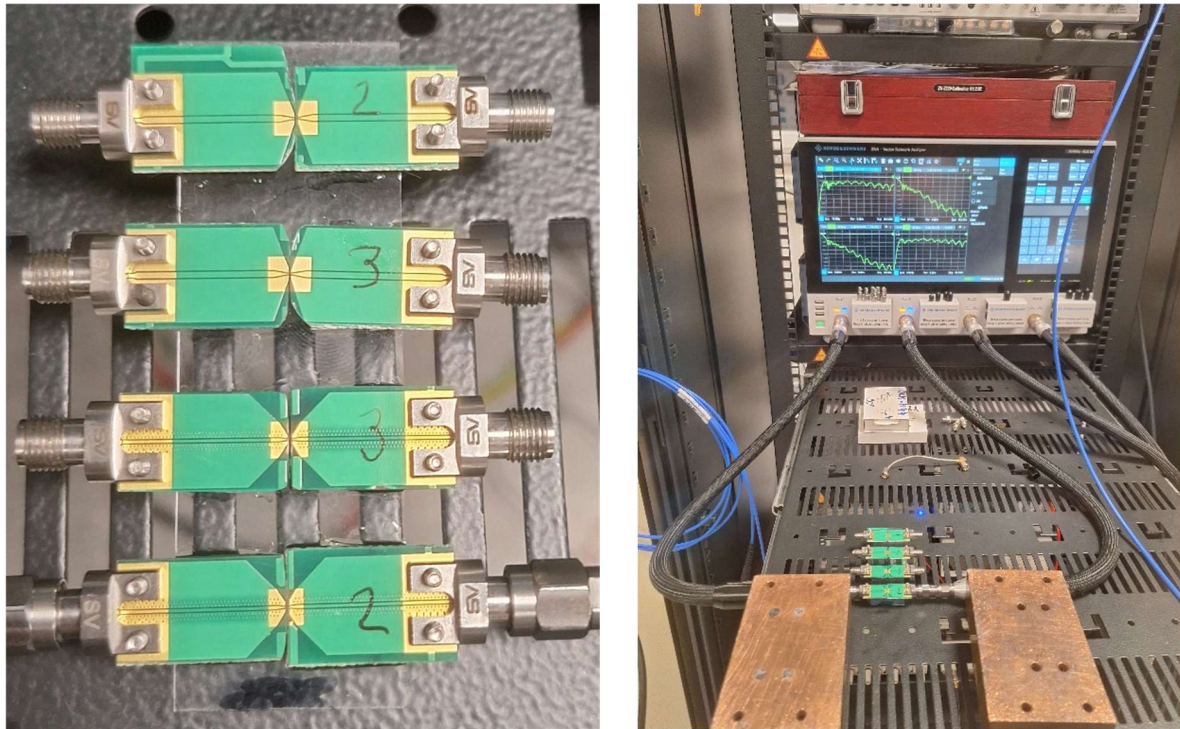


Figure 4.25: Left: A sample to measure the S -parameters of the ring modulator fan-out boards and to characterize the penalty of wire bonding only one grounded plane as opposed to both. Right: The setup used for the characterization.

Figure 4.26 and 4.27 show the results of these measurements. Due to the symmetry of the sample pieces $S_{12} = S_{21}$ and $S_{11} = S_{22}$ can be postulated within measurement uncertainty, thereby only one of each needs to be discussed. The S -parameters only allow judgments of the whole sample and not for the individual PCB or wire bonds. All S_{12} follow a similar linear path from 10 MHz to 43.5 GHz, starting at roughly 0 dBm and ending at -20 dB, but with some ripple. At 39 GHz the CPWs have a steeper fall-off to approximately -27 dB. A steep, noisy drop in the two wire bond CPW curve around 2 GHz can be observed. The S_{11} 's starts in comparison to the rest of the curves at a very small value, then at approximately 6 GHz settling in the -5 to -15 dB range. The two wire bond CPW design reaches this range before 2 GHz but then starts oscillating until a dip at around 5.5 GHz, after which it settles with like the others. All curves show regular oscillations until -43.5 GHz in addition to the increasingly worse magnitude. Both CPW samples show a stronger fall-off at frequencies higher than 39 GHz.

While discussing these curves, we need to remember that the back-to-back configuration of the fan-out PCBs is a worst-case scenario. In the intended configuration in section 5.3 only one is used. Therefore, the linear decay of the S_{12} , which shows the transmission losses, would flatten possibly by up to 50 %. For such a wide band application, this is acceptable. Regarding the difference between the CPWs and GCPWs, clearly the GCPW designs prove to be superior. Especially considering the strong resonance for the two wire bond CPW at 2 GHz and the general CPW decline over 39 GHz. The former clarifies the advantage of GCPW for dealing with open-ended ground planes. The vias and connected internal copper layers enable to draw the non-bonded plane to match the other sides' potential. The stronger decline at higher frequencies is most likely not only caused by the potential mismatch but also leakage due to the inferior field confinement and lower total inductance of the design. High frequency currents take the path of lowest inductance and the greater surface of the GCPW design tends to lower it. To

increase the performance of the CPW two wire bond sample, it would possibly help to wire bond the two ground planes together close to the taper tip.

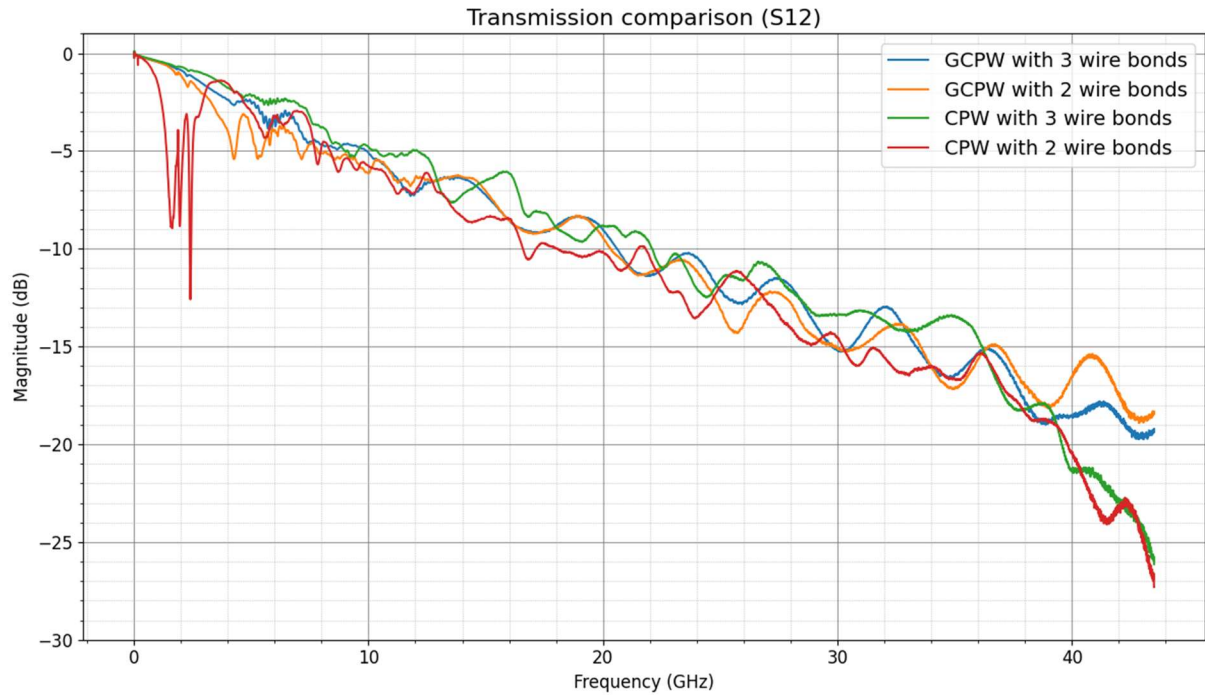


Figure 4.26: Transmission comparison between the ring modulator fan-out boards and the amount of wire bonds used.

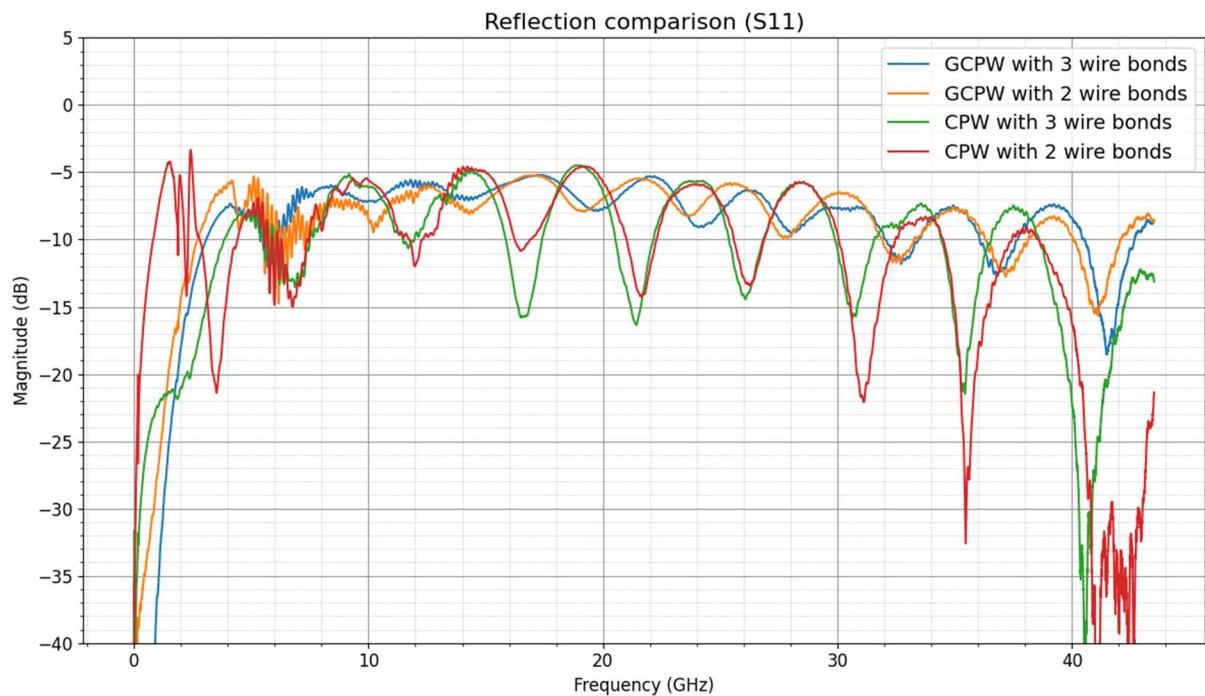


Figure 4.27: Reflection comparison between the ring modulator fan-out boards and the amount of wire bonds used.

Looking at the S_{11} shows that especially the GCPW designs are prone to parasitic resonances around between 5 GHz and 8 GHz, likely due to substrate-mode coupling (parasitic modes between the surface ground planes and the buried ground plane [45]). In contrast, the two wire bond CPW variants exhibit an apparent improvement slightly above 40 GHz, which is misleading and an artifact caused by energy radiation (antenna effect/electric field leakage) rather than superior matching, as confirmed by its poor transmission observed in the S_{12} . Despite the ripples and on average higher decibel values in the S_{11} , the GCPW remains the better choice, as it ensures the more stable field confinement, higher overall power transfer and predictable behavior over the full frequency range. As expected, three wire bonds achieve better characteristics, but the GCPW design compensates most of the penalty.

Performing time domain reflectometry, which would show, where along the RF signal path the impedance matching failed, and a time domain transmission graph, which would show the signal warping and dampening, was attempted. However, no plots with a conclusive picture could be created.

Until the full integration advantage, meaning driver electronics on a chip, next to or even on the photonic chip can be harnessed, future samples can further be improved. Apart from the choice of RF dielectrics for the fan-out PCBs, specialized RF wire bond technologies should be implemented. These include industry standard ribbon bonds or wire bond stitching, where several wire bonds connect each pad with the corresponding surface on the PCB. If GCPW or CPW should be kept, a three-pad design on the PIC would be advantageous. However, the CPW design should be dropped completely.

4.4.2 Fan-out PCBs for Mach-Zehnder modulators

The fan-out PCB requirements for the MZM designs are more complex compared to the RM ones. The first change is that the MZM needs to be connected from two sides via fan-out boards. Due to the MZM design, it cannot serve as the termination resistor as the RM could previously. Thus, a dedicated termination resistor must be placed at the end of the signal chain. It is mounted on that secondary fan-out PCB. This additional PCB will be referred to as termination fan-out PCB. The sinusoidal RF input signal in the experiments in chapter 5 will be driven with 0 dBm input power. Termination resistors, which are certified up to 40 GHz or further, are very small. They cannot effectively dissipate this amount of power. Therefore, an RF attenuator is placed in front of the termination resistor in the signal path. The layout of the termination fan-out PCB is shown in figure 4.28. The large pads are there to solder a two-pin connector on, so the potential can be set to ground to avoid a floating potential and their possible negative consequences. The dense amount of vias in the middle are there to cool the RF attenuator properly.

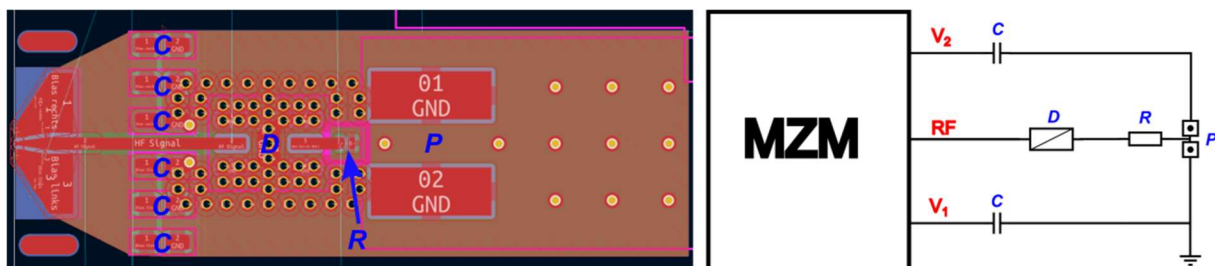


Figure 4.28: Left: Grounded coplanar waveguide layout of the termination fan-out PCB. Right: Circuit diagram of the same layout.

Designation in figure 4.28	Component type	Component name
D	3 dB RF attenuator	Mini-Circuits YAT-3A+
R	50 Ω termination resistor	Vishay CHA02016-50RGFTD
C	1 μ F capacitor	Kyocera AVX 550Z104MTT
P	Connection pins	Samtec TSM-102-01-T-SH

Table 4.4: Components of the termination fan-out PCB as of the circuit diagram in figure 4.28.

The push-pull design of the MZMs on the COTTONTAIL chip sets the second additional design requirement. The GCPW or CPW outer surface planes around the center RF conductor need to be biased in opposing directions compared to the latter. This needs to happen on both sides, i.e. both PCBs. This is realized by a capacitive decoupling. The surface planes are interrupted and therefore isolated for DC currents. However, through capacitors (preferably as close as possible to the inner RF conductor) the RF return currents can still pass through the outer planes. This design comes with many expected penalties. Both the CPW and GCPW now have different potentials on all planes. For GCPW also the buried plane has different potential, i.e. it is on ground. This necessitates disadvantageous RF mode transformations. Via fencing is not possible anymore, where the potential varies.

The other change due to the push-pull MZMs is the need for DC supply arms, not unlike bias-tees. Here it doesn't act as a combiner of a DC and RF signal. The bias-tee acts as an RF barrier, blocking parasitic RF signals from leaking into the DC path. The R-C connections to ground help with the attenuation of the remaining parasitic signals (de-Q), following the general bias-tee introduction in section 2.3.

One GCPW layout example with such bias-tees is shown in figure 4.29. The design was adapted from the AN-2061 wideband bias-tee [37]. The design there was tested up to 30 GHz with promising results. For any layout presented in this thesis two bias-tee arms are necessary, since two unperturbed DC sources are needed. These arms can be seen on the outer sides (the very left and very right) of the layout connected to the outer pad of the large pin footprints marked with 01 and 02. The corresponding pin supplies the DC bias. The inner pin is an additional ground to give any parasitics a low impedance path away from the MZM. The conductor lines towards the bottom planes of the taper, where the wire bonds will be attached, carry the DC signals. The paths leading away from the DC lines are the R-C pairs.

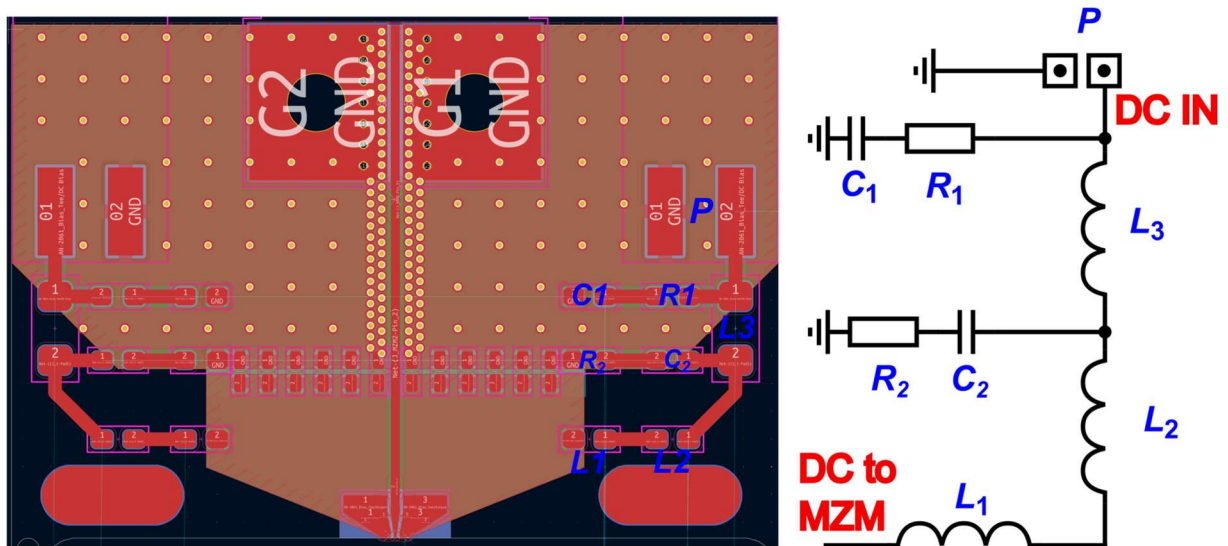


Figure 4.29: Left: Layout of the AN-2061 inspired DC-Part of the bias-tee [37]. Right: Circuit diagram of one bias-tee arm.

Designation in figure 4.29	Component type	Component name
R1	0 Ω jumper	Vishay CRCW04020000Z0EDHP
C1	10 μ F capacitor	TDK C1005X7S2A103K050BB
L3	470 nH inductor (ferrite core)	Coilcraft 0805LS-471XJRC
C2	100 pF capacitor	Yageo CC0402JRNPO9BN101
R2	340 Ω resistor	Panasonic ERA-2AEB3400X
L2	56 nH inductor (ferrite core)	Coilcraft 0402DF-560XJRW
L1	20 nH inductor (ferrite core)	Coilcraft 0402DF-200XJRW
P	Connection pins	Samtec TSM-102-01-T-SH

Table 4.5: Components used for the bias-tee shown in figure 4.29.

Any MZM sample for the irradiation measurements discussed in section 5.4 will be connected like shown in figure 4.30. The capacitor separating all DC and ground planes is the Kyocera AX 550Z104MTT (to form a complete bias-tee this capacitor should be classified as part of it). The RF connector on all layouts is placed on the large footprints centrally connecting the center conductor. All conceived layouts are briefly introduced in the Supporting materials chapter at the end of this thesis, since they weren't used in any measurement.

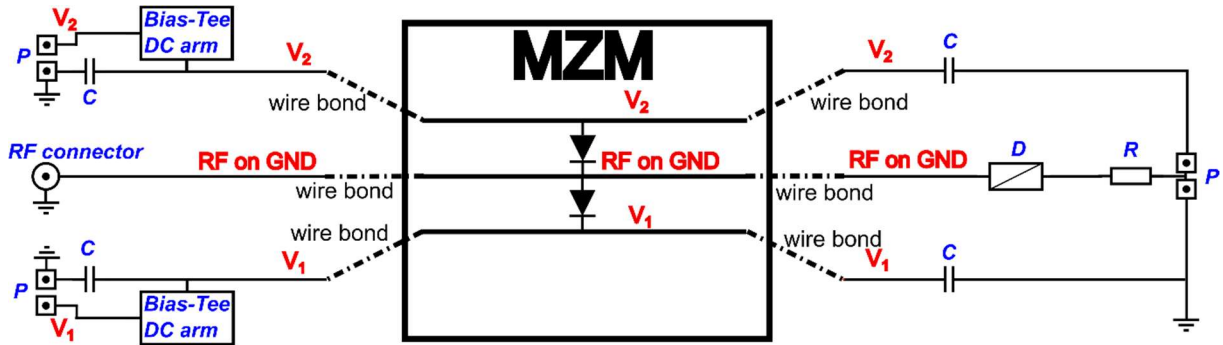


Figure 4.30: Circuit diagram of a full Mach-Zehnder modulator sample. The left side is the fan-out PCB holding the bias-tees and RF connector. In the center is the Mach-Zehnder modulator. The pn-junctions within are approximated by diodes. On the right is the termination fan-out PCB. All individual parts are connected via wire bonds. To ensure a reverse biasing of the modulator and therefore the push-pull functionality $V_1 > \text{GND} > V_2$ must be true.

Due to time constraints no RF characterization measurements were conducted for any MZM design.

4.5 Sample builds for irradiation measurements

This chapter shows the culmination of all the preparational work by combining the findings into complete sample builds. Here, the method is introduced, whereas the subsection will show the modifications due to arising issues to this process and their specifics for the MZMs and RMs.

The process starts with the fiber preparations as shown in section 4.2. The fibers are then glued into a fiber holder/handler like the one shown on the right in figure 4.31. In this step the overhang of the fiber and its rotational alignment is determined. The goal is an overhang of 5 mm. For this a newly conceived setup will be presented in the following subsection. The fiber holder is 3D printed with an Anycubic Photon M3 Premium resin printer and Phrozen Aqua Gray 8k resin using a 10 μm layer thickness. The glue used for fixing the angle-polished fiber to the fiber holder is the previously mentioned OP-54. The fiber holders come with a groove for a wedge that helps fixate the fiber before gluing and increases the bonding surface between fiber holder and fiber. The fiber holder is asymmetrical to provide a large uniform surface area to be handled by a micrometer screw driven vacuum pickup.

Due to space constraints and economic reasons (otherwise a whole chip is needed for one sample) the COTTONTAIL chips are diced on a Disco DAD3240 wafer saw. The result of the dicing is shown in figure 4.31 (left). The chips are then cleaned with adhesive tape. It is attached to the surface of the diced chip and pulled off again, removing most residues from the diced chip's surface. This was repeated two more times with fresh tape. After this it is cleaned further with acetone and finally isopropyl alcohol.

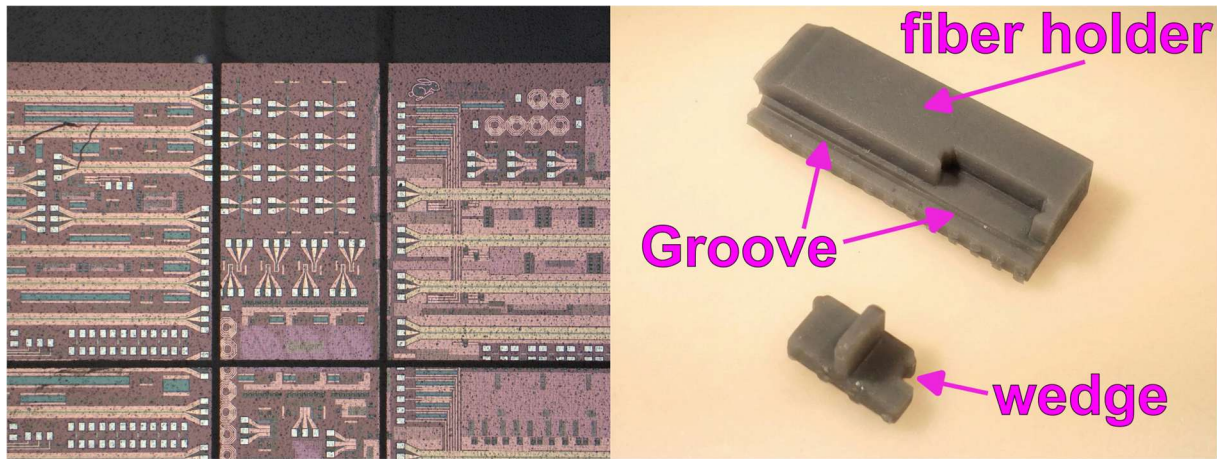


Figure 4.31: Left: Diced COTTONTAIL chip with the ring modulators in the top center. Right: 3D printed fiber holder and wedge.

The samples are built with a standard microscope slide as a base, which is placed on the vacuum stage. The part of the diced chip containing either the MZMs or the RMs is placed on the slide on top of some OP-54 glue, so that the GCs are oriented parallel to the long side of the slide. The slide must be positioned with enough overhang, so that the chip bottom can be reached with a UV source to harden the glue. Overhang is necessary for every part that needs to be glued. This specific orientation of the diced chip is necessary for two reasons: First the fiber coupling needs enough space to glue down the fiber holders with the fiber. Second, since a minimal distance between structure and RF connector is advantageous, the fan-out PCBs could be shortened. As hinted at in subsection 4.4.1 this together with the necessary overhang to facilitate the use of the RF connector chosen, was the design constraint to the fan-out PCBs length.

Next the fan-out PCB(s) wire bond sides are ground down close to the taper layout shape from figure 4.21. This has two reasons: mainly to increase the space available to the fiber holders, but also to minimize glue creep towards the chip surface. The latter needs to be considered for PCB mounting on the slide. Optimally the diced chip and PCB would be mounted as close to each other as possible to shorten the wire bonds. However, the glue creep necessitates a larger minimal distance, to prevent glue on the wire bond pads and GC. After the glue was hardened just from underneath and through the microscope slide, the chip surface is wiped with acetone followed by isopropanol, without drenching the sample. After this cleaning step, the glue was exposed to UV light from all sides. Now the wire bonds can be set.

Finally, the angle-polished fibers with their holders can be mounted. A sketch of the proposed mounting system is shown in figure 4.32. This is done in the DC setup introduced in section 4.1. The output side remains in the flat-faceted fiber configuration, while the input side uses the vacuum pickup like shown in figure 4.40 with a PM fiber. When the coupling is optimized, the fiber is glued down with the OP-54. Ideally the glued fiber is suspended above the chip surface without coming into contact with it. The fiber should now be optimally placed, so that the sample can be rotated and the output fiber in the fiber holder can be glued in place as well. Then, the RF connector is screwed onto the fan-out PCB with the corresponding footprint, which results in the complete sample for the irradiation measurement. The electrical connections are tested with a multimeter. To ensure a flat and rigid assembly after every gluing step excess glue is removed with a scalpel. If the sample has a lot of access glue on it, it needs to be

shortly drenched in acetone first. For stability the fibers are also glued to the edge of the microscope slide.

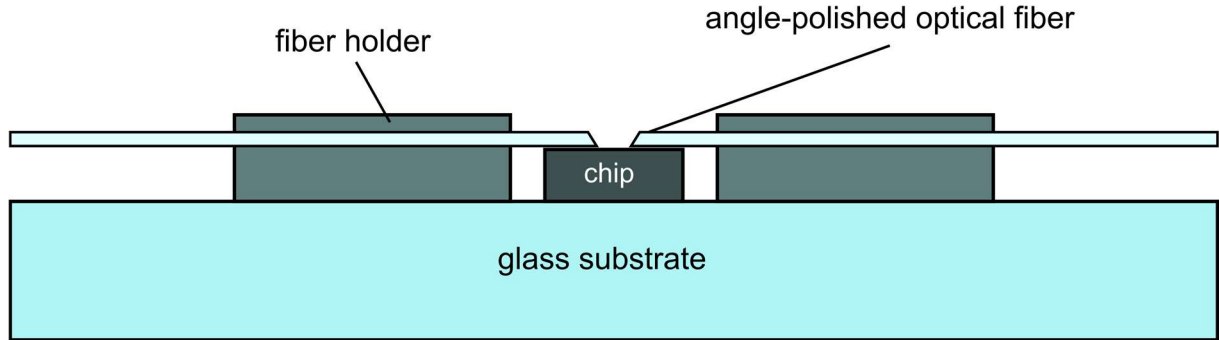


Figure 4.32: Sketch of the arrangement of a full sample's parts using flat fiber-to-chip coupling.

4.5.1 Setup for rotational fiber alignment

As part of this thesis a setup for the rotational alignment of the fiber in the fiber holder has been developed. Previously the alignment was done on the DC setup from section 4.1 above the device under test. The coupling was optimized, then the angle-polished fiber was moved away while still being clamped in the fiber rotator shown in figure 4.4 and then placed into the fiber holder. This method works but doesn't allow for the quantification of the rotational alignment and is very slow. The difficulty in accurately rotationally aligning is further increased by oscillations in the optical power received by the OPM. The new setup therefore helps with aligning the fiber in the fiber holder, without using the main setup and a chip and allows for a statement about the rotational alignment accuracy.

Its basic idea is that the angle-polished fiber is connected to a laser and rotationally aligned over a photodiode while already being placed but not glued to the fiber holder. Then the photocurrent should be maximized to reach optimal rotational alignment.

Beforehand some preparational calculations were done to check the feasibility of such a setup. This can be determined by the optical power arriving at the photodiode. This depends on the input optical power, the distance between fiber and photodiode and the photodiode diameter. Increased distance in turn improves the accuracy of the alignment.

To make a judgment of the projection of the optical power distribution on the photodiode surface needs to be calculated. The problem can be split into two 2D planes. The first one is perpendicular to the fiber (which means following the normal propagation direction of the fiber) after the TIR on the angle-polished facet. The two beam waists follow a radial propagation path out of the fiber core. This reflects the behavior after z_R as introduced in 2.1.2.

The second 2D plane is parallel to the fiber (view at the side profile of the fiber above the projection plane). Here the propagation becomes a more complex problem to solve. The center of the beam and both waists need to be tracked. Within the fiber the waists still diverge since $r_{r, \text{fiber}} < z_R$. Therefore, both beam waists and the beam center hit the bottom of the fiber at a different angle therefore causing Snell's law to warp the distribution further. The Gaussian distribution is therefore warped twice by the widening of its waist. A schematic of this is shown in figure 4.33. To visualize the calculation, it is easy

to show that after the TIR without diffraction the Gaussian beam in the fiber core would simply be reflected to the bottom of the fiber. But diffraction exists, so the Gaussian distribution must be stretched. This can be calculated with equation (2.10) and simple trigonometry. The new angles with which the waists hit the fiber bottom also can be derived. Then Snell's law is applied to these new angles. Then the waists are followed with the assumption of no further diffraction. Finally, on the projection surface i.e. photodiode, which's normal vector is assumed to point towards the fiber, further stretching needs to be applied to the already warped distribution. This depends on the distance between the fiber bottom and the photodiode. The result is a continuous distribution which needs to be normalized to the original input power. The resulting distribution for a 1550 nm laser with a power of 0 dBm at a distance of 4 cm is shown in the heatmap in figure 4.34.

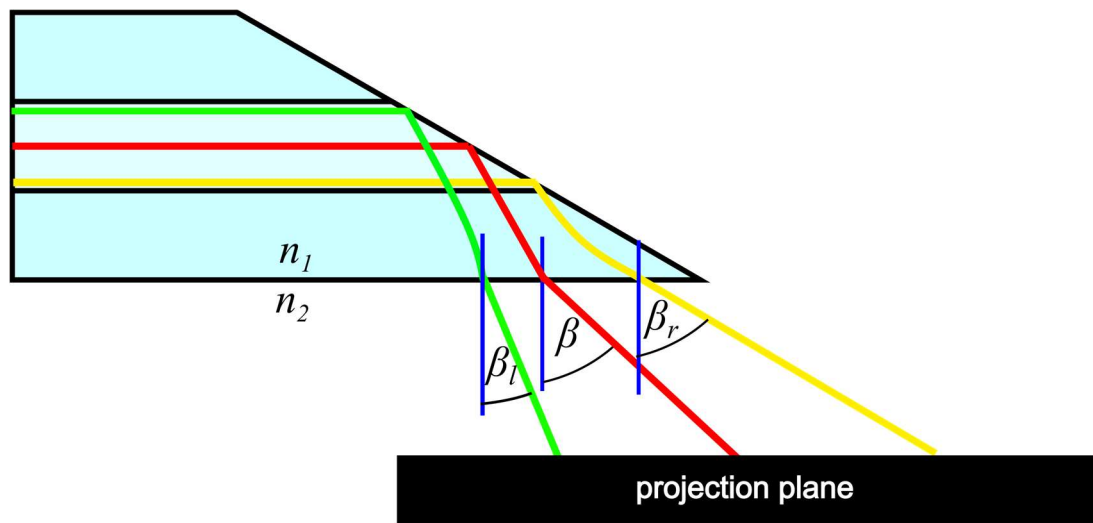


Figure 4.33: Widening of the Gaussian beam after the reflection at the angle-polished facet, due to diffraction. This results in different refraction angles on the fiber bottom for the beam waists and the beam center.

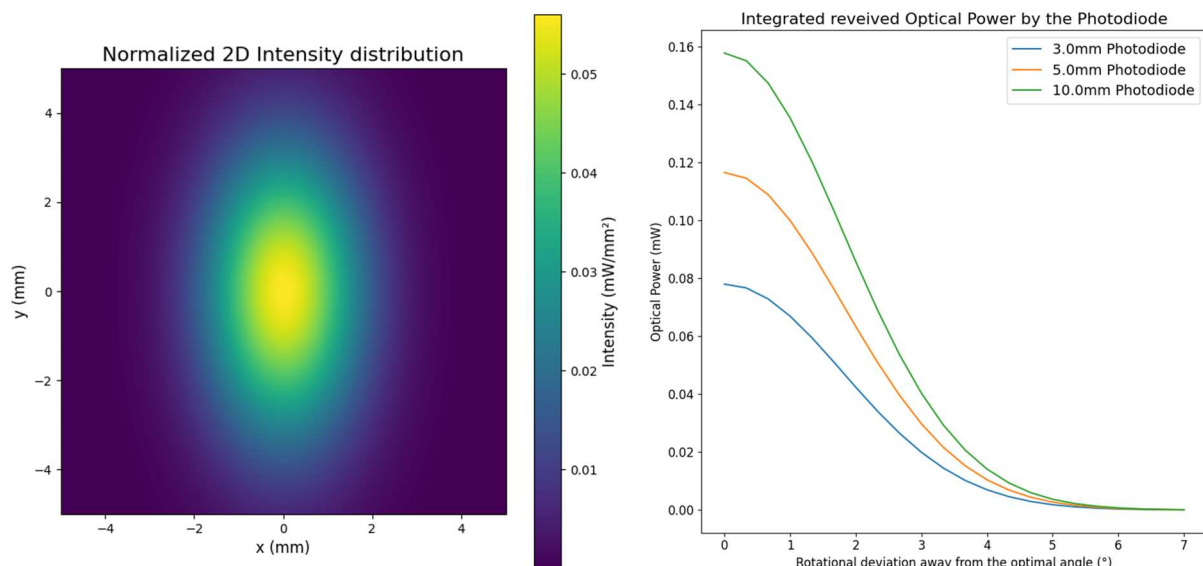


Figure 4.34: Left: Intensity distribution on a 2D projection plane 4 cm away from the fiber bottom as arranged in figure 4.33. Right: Approximated received optical power by photodiodes with different diameters at the same distance through a 0.5 mm slit aperture. The assumed laser output power for both is 0 dBm.

To increase the sensitivity to rotational misalignment an aperture is placed above the photodiode. The best estimate for a reliable aperture slit width printable on the resin printer introduced in section 4.5 is $500\text{ }\mu\text{m}$. Finally, a curve could be drawn, how the received optical power changes with different photodiode diameters at angles away from the optimal angle and more importantly, if this power can be detected. The result of this is shown on the right in figure 4.34. The rotational change is approximated by simply shifting the aperture windows along the heatmap. The estimates for the received optical power therefore reach from roughly $80\text{ }\mu\text{W}$ to $160\text{ }\mu\text{W}$. The detectability depends on the responsivity $R(\lambda)$, the noise of the photodiode and the measurement sensitivity of the attached instrument [46].

To jump ahead the FDG03 Ge photodiode (Thorlabs Inc.) with an active area diameter of 3 mm was employed. It has a responsivity of $R(1550\text{ nm}) = 0.85\text{ A}\cdot\text{W}^{-1}$. This leads to a peak expected photocurrent of approximately $68\text{ }\mu\text{A}$ for a laser output power of 0 dBm, which is easily measurable with instruments like SMUs or many digital multimeters (DMMs).

Therefore, the feasibility of the conceived setup could be proven. The finalized setup is shown in figure 4.35. It incorporates a newly designed, FDM printed stage, which holds the photodiode and has spaces for two resin printed inserts, one to fix the fiber holders and the other one is the aperture. It also holds the in [46] recommended circuit to run the photodiode shown in diagram figure 4.36. Two BNC connectors are attached to the circuit, one which supplies the voltage and one to read the voltage drop over R_L . The latter would be used to calculate the photocurrent. This was later changed to a digital multimeter reading the current input via the voltage supply, making the second BNC port obsolete. Thus, a new circuit was made dropping R_L . The fully equipped stage with the rotator stage, which holds the fiber rotator with the micrometer screws and angle scale, is shown in figure 4.37.

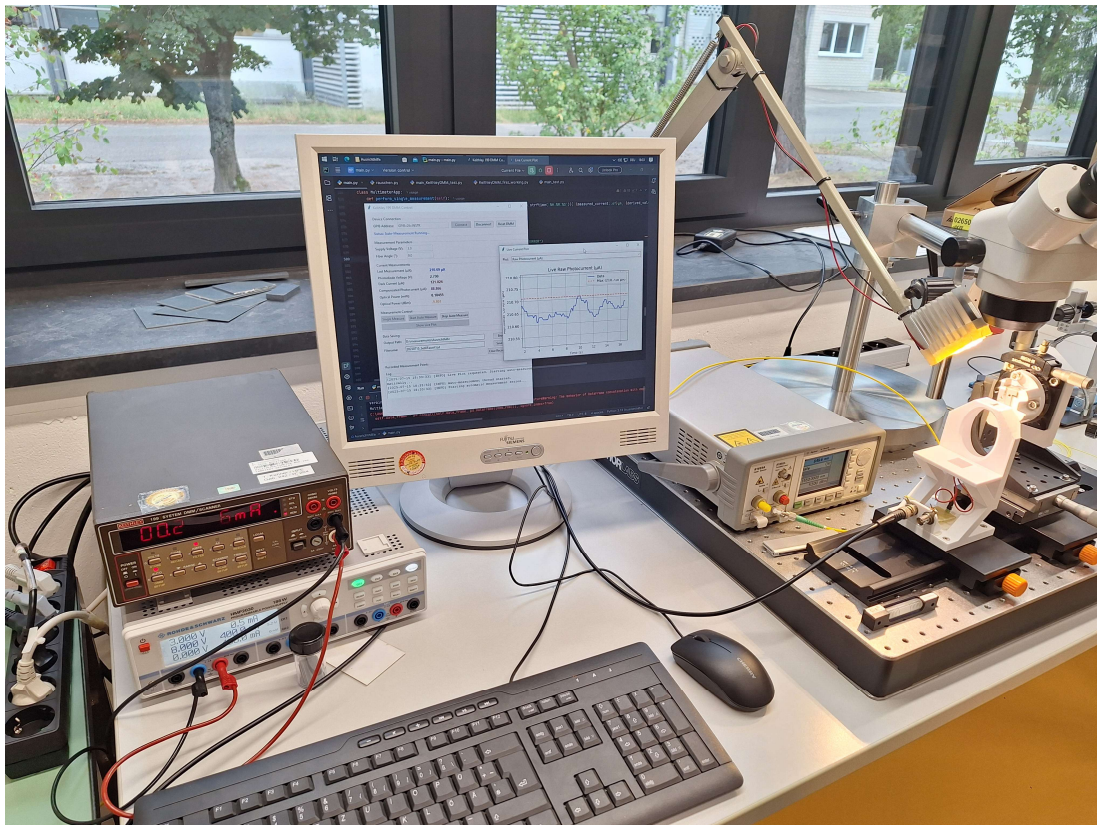


Figure 4.35: The new setup to rotationally align angle-polished glass fibers within the fiber holder for full sample builds.

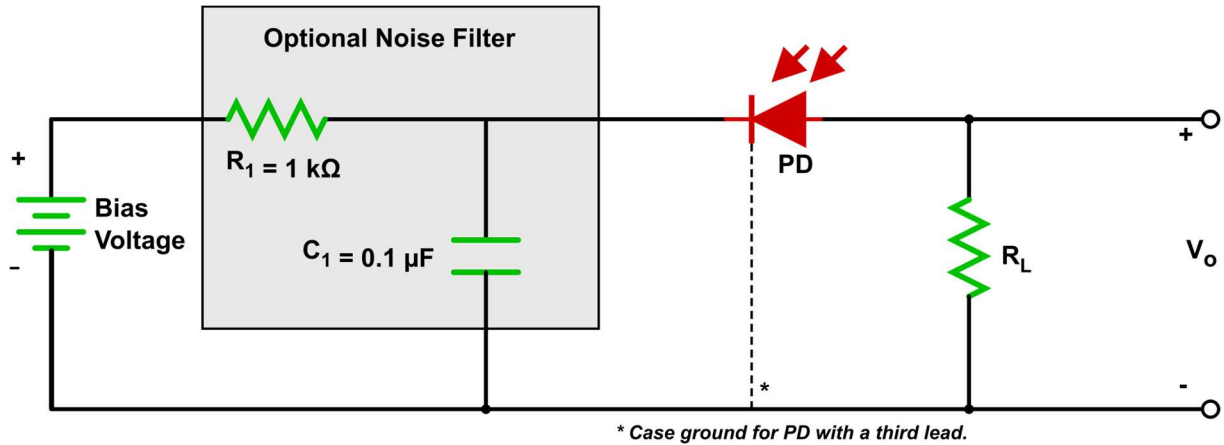


Figure 4.36: Recommended circuit do drive the FDG03 photodiode (adapted from [46]).

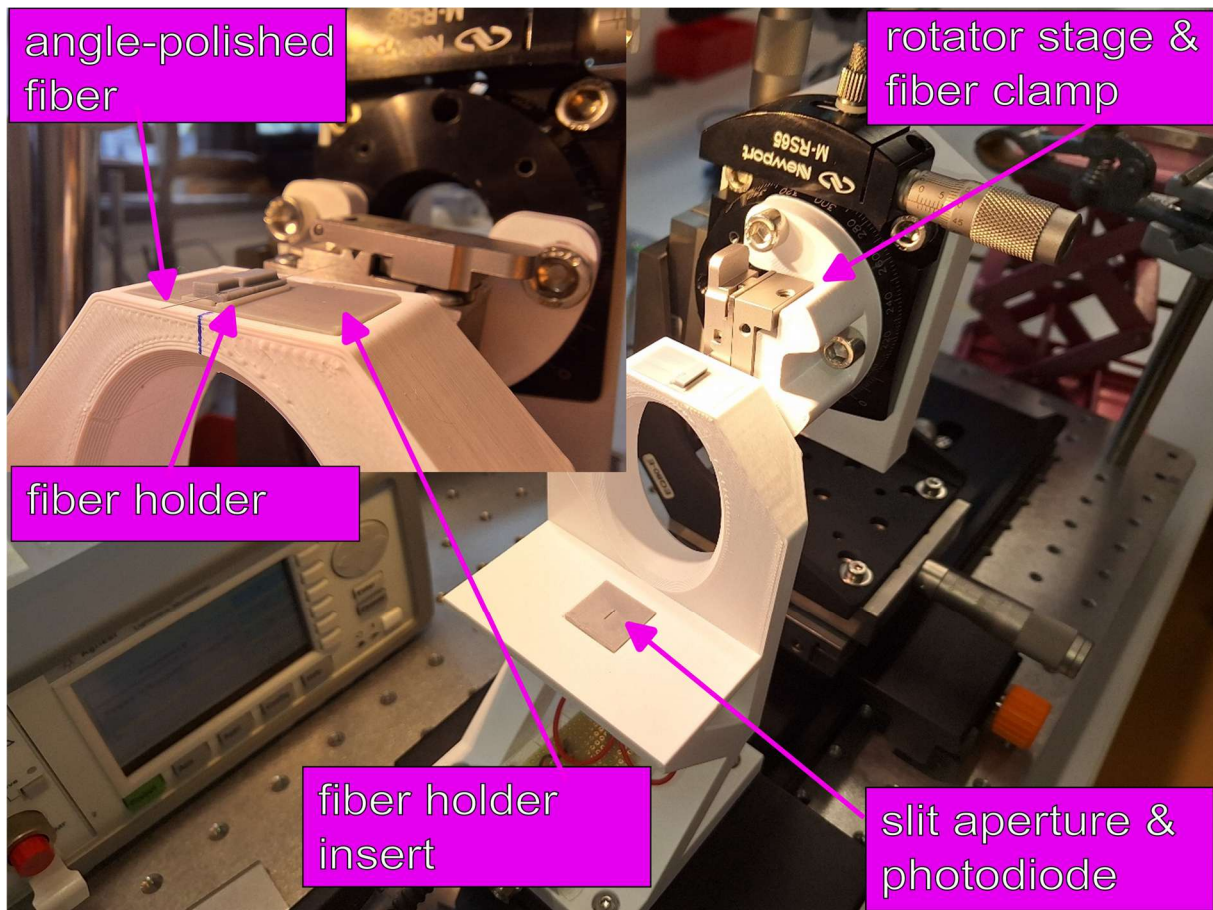


Figure 4.37: Close-up image of the alignment stage of the setup.

The external devices used are a HMP2030 programmable power supply (Rohde & Schwarz), a Model 199 System DMM (Keithley Instruments) and a 81950A Tunable Laser Source module in a 8163B Lightwave Multimeter (both Agilent Technologies). Note that the minimum suppliable laser power of the laser module is 5.5 dBm, which leads to higher photocurrents compared to the 68 μA estimate. The DMM has an internal resistance which needs to be considered in the circuit diagram.

Next all resistive components were measured to their exact value. A driver for communication with the DMM via GPIB with basic functionality was coded in Python with the aid of [47]. Also, a dark current calibration curve with respect to the voltage over the photodiode was taken. A polynomial fit for the resulting curve from this measurement was added to the main program of the rotational alignment setup. This enables the calculation of the true photocurrent, if necessary. The main program also manages the driver integration, allows single data point measurements with a manually set input angle, or the recording of real time data. The real time data can simultaneously be plotted in a rolling plot.

The display for real time data is the function used when the setup is utilized for aligning and gluing fibers. The supply voltage chosen is always 3 V but can be set to any value. First the fiber is connected to the laser and fixed in the fiber rotator. Then under a microscope the fiber is positioned above the fiber holder and then lowered into its groove. A rough alignment of the angle-polished facet is done, so that the light hits the surface of the rotator stage, where the aperture and photodiode are located. Then the laser output is enabled. Via the live plot of the photocurrent the output is maximized via rotational and longitudinal alignment. The latter is possible since the geometry of the fiber stage is such that if the fiber has a correct polishing angle, the maximum of the measured quantity is reached with the desired 5 mm fiber overhang.

To evaluate the rotational alignment accuracy, two measurements are required: the photocurrent as a function of the angle, and the system noise at optimal alignment. The noise level is critical because it inherently limits the precision with which the optimal rotational angle can be determined.

The measurement results of the first characterization step are shown as blue data points in figure 4.38. These points are then fitted with a Gaussian distribution (red). To quantify the alignment precision limit, the photocurrent was recorded over several hours at a fixed optimal position using the maximum sampling rate of the DMM. The resulting data was smoothed using a Savitzky-Golay filter (using a third order polynomial and a 101-point window) to establish a baseline. The residuals (deviation of raw data from the filtered curve) were then plotted in a histogram and fitted with a Gaussian distribution.

Assuming that the alignment is only distinguishable within a confidence interval of three standard deviations (3σ), the noise distribution effectively blurs the precise location of the peak. As shown in figure 4.38, the two vertical green lines indicate the angular range corresponding to this noise floor. This results in an alignment accuracy of $\pm 0.1^\circ$.

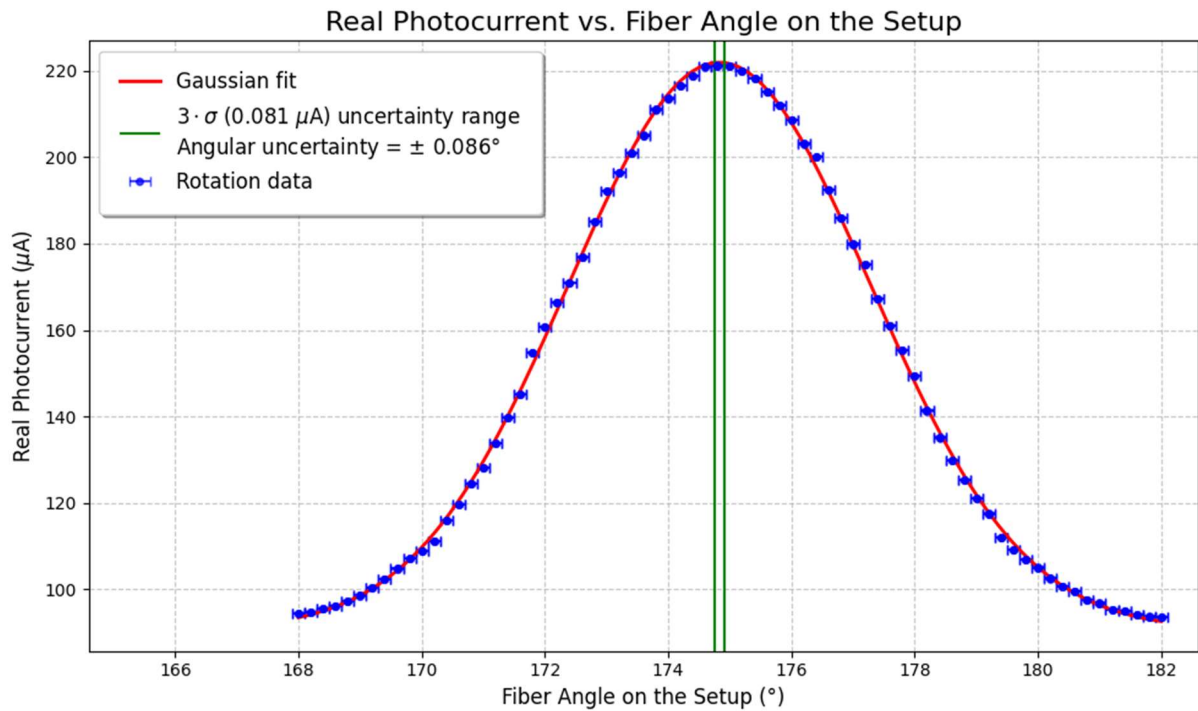


Figure 4.38: Measured photocurrent plotted against the rotational angle of an angle-polished fiber within the setup with error margins.

4.5.2 Ring modulator samples

Figures 4.39 to 4.41 show the different stages of a RM sample built. The general process was described in section 4.5. However, a modification was made: The plastic resin 3D printed fiber holders as shown on the right in figure 4.31 were swapped with ones made of glass. First irradiation measurements have shown that the resin printed fiber holders and the glue warp under X-ray irradiation. Therefore, optical coupling was either lost immediately, or very quickly, before any radiation damage to the chip could be observed. This can be seen in figure 4.42.

Since the fiber holders were determined to be the main cause for this behavior, glass printed replacements were deployed. They originate from an ongoing collaboration with S. Liu (PhD cand.) from the Institute for Micro Process Engineering of the KIT and are printed in a similar manner as described in [48]. However, the dimensions of these glass-printed holders are not easily controlled. The final design used is too tall, therefore decreasing the optical efficiency. However, to a more limited extent this has also been noticed with the plastic resin printed fiber holders. Generally, for the irradiation measurements for the proposed optical interconnect with the floating fibers over the chip surface, the glass-printed holders provide the required radiation hardness. Due to time constraints, no insert fitting to glass-printed fiber holders for the rotational alignment setup was printed. A simple flat surface was used, most likely decreasing the error margin of the rotational alignment accuracy.

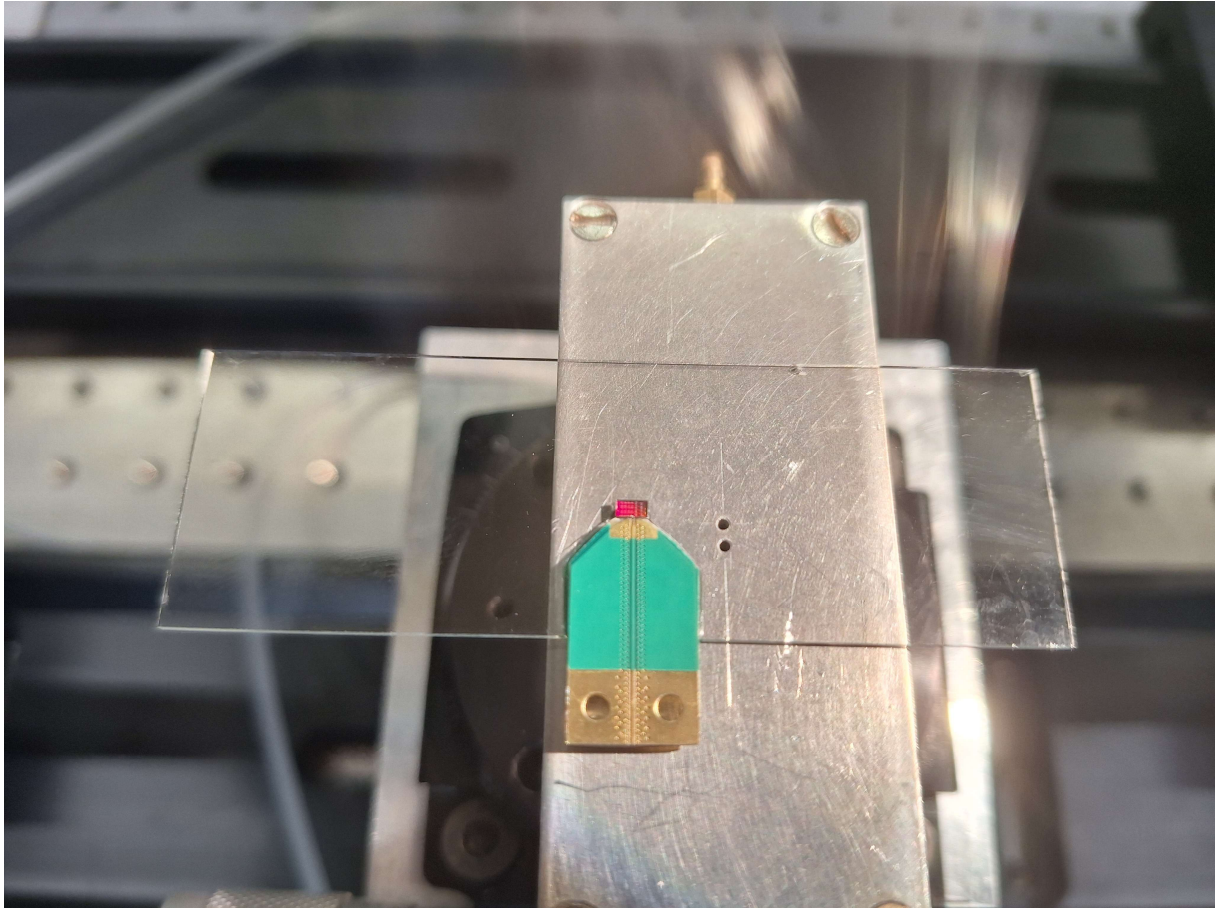


Figure 4.39: First stage of the ring modulator sample build. The diced part of the chip, which contains the ring modulators, was glued down first. Then the sample was cleaned and the fan-out connection PCB was glued close to the connection pads. A gap between both parts was left to combat glue creep.

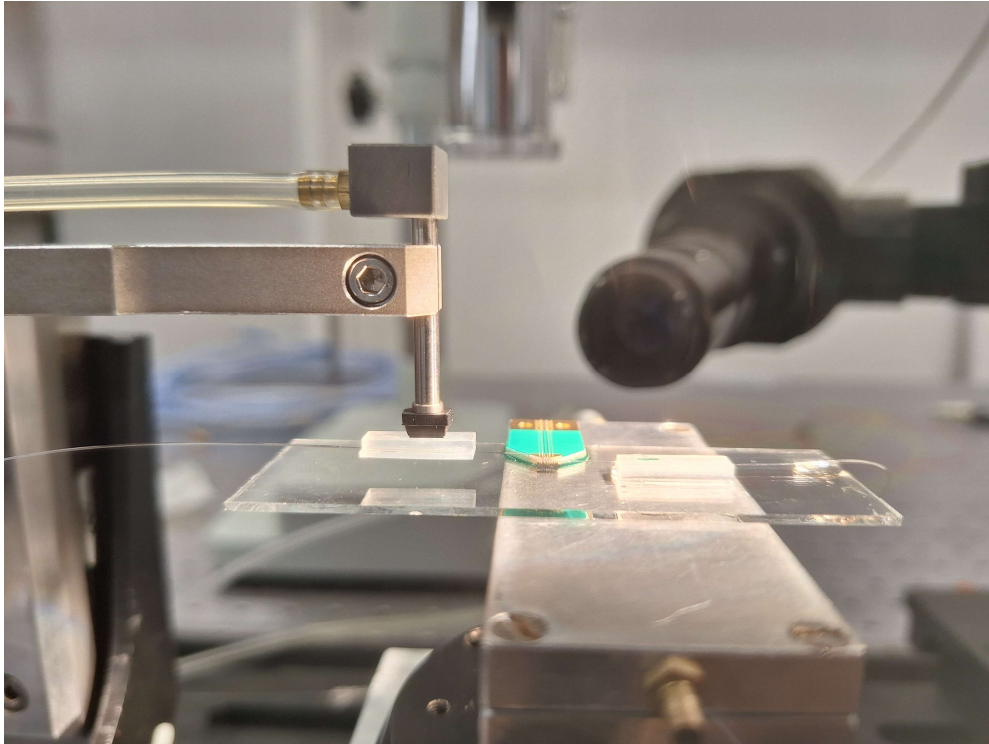


Figure 4.40: Second stage of the sample build. The fiber holders (here the glass versions) containing the angle-polished fibers are handled by a vacuum pickup and the coupling is optimized. Then the fiber holders are glued down.

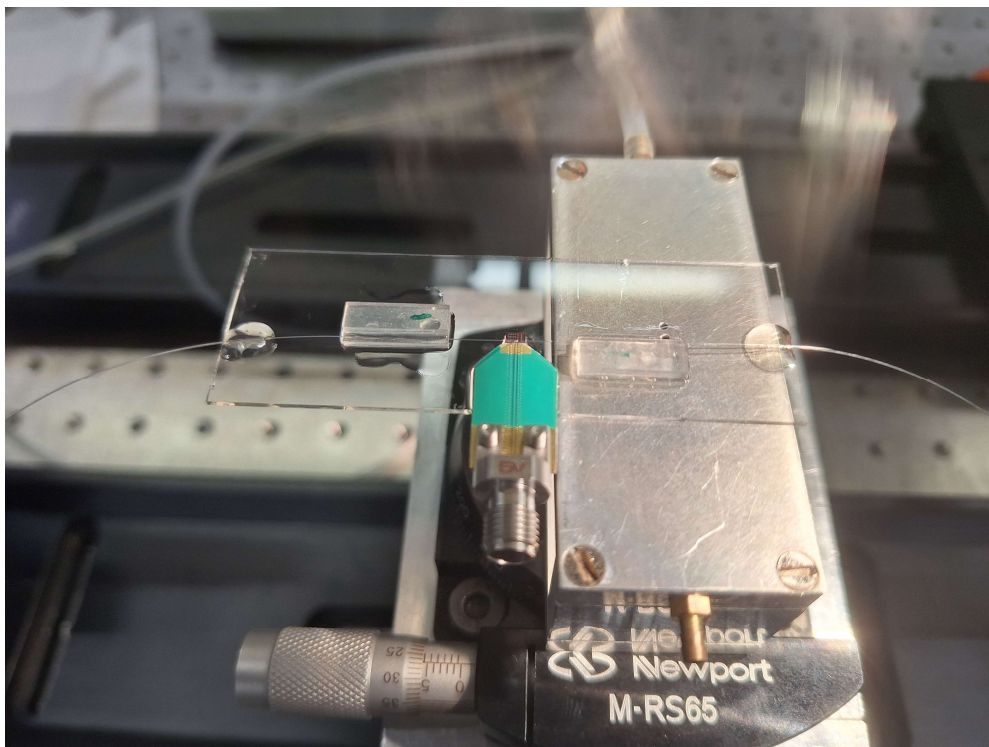


Figure 4.41: Fully assembled sample. The fibers were glued down on the side of the microscope slide to increase rigidity. As a final step the RF connector was screwed down on its footprint.

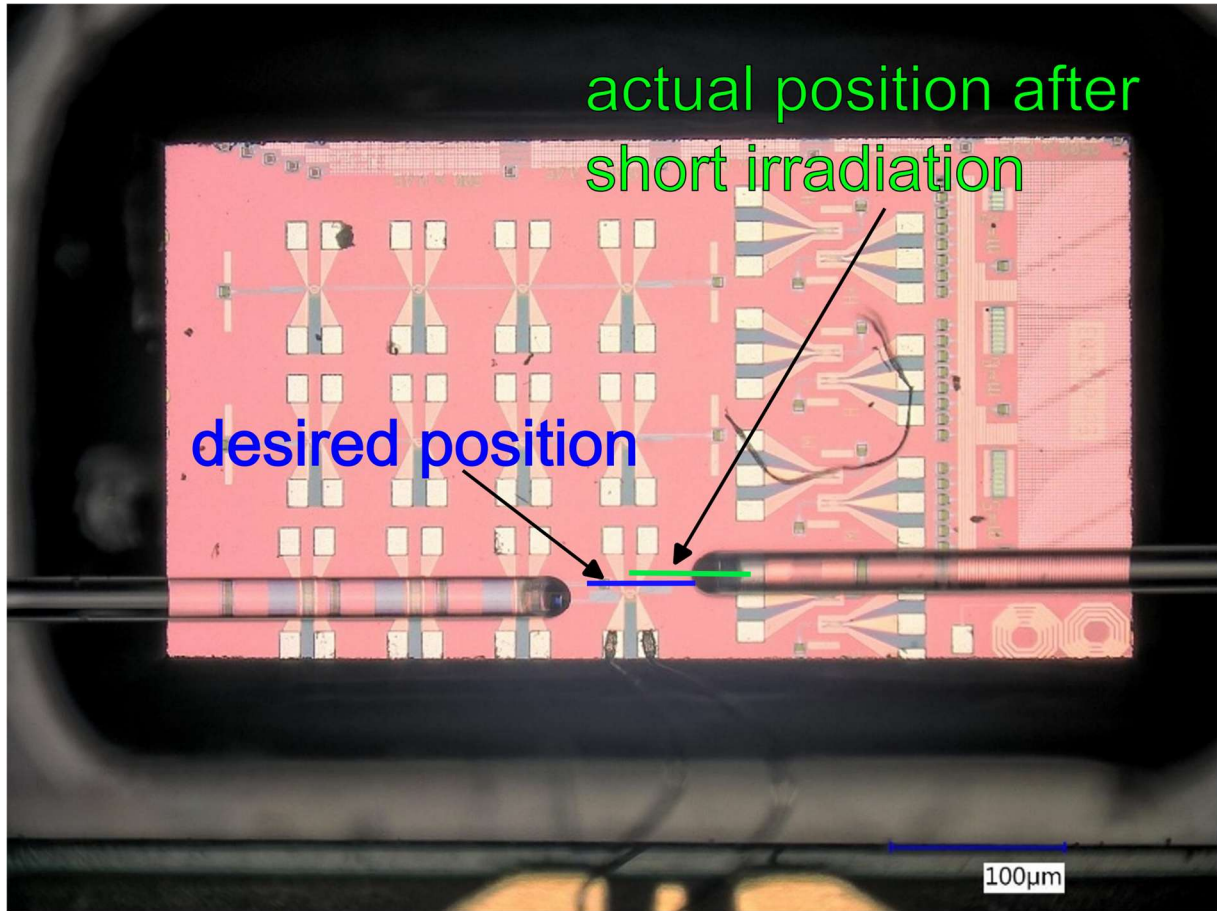


Figure 4.42: Angle-polished fibers held by plastic resin 3D printed fiber holders. The left fiber is still above the grating coupler. The right fiber shifted towards the top of the image (only the right fiber holder was irradiated by the primary X-ray beam due to bad alignment).

For a non-irradiated RM GCPW (two wire bonds) sample, S_{11} parameters were recorded. This was done with the same setup mentioned in subsection 4.4.1. Since the RM sample electrically is a single-port device, only a S_{11} can be recorded. The resulting curve can be seen in figure 4.43. Up to 15 GHz a mostly linear decrease in the curve from 0 dB to -7 dB can be observed. After that point an oscillating behavior up to 43.5 GHz is dominant. Together this can be interpreted as a transmission dominant behavior until 15 GHz. After that, the transmission component i.e. the power of the modulation going through the RM vanishes in the reflections and cannot be easily foreseen. The reasons for the exact behavior are the same as mentioned in subsection 4.4.1. The oscillations are most likely caused by the wire bonds. The irradiation measurements will make the transmission and reflection component more distinguishable, since the EOE response is mainly reliant on the transmission component.

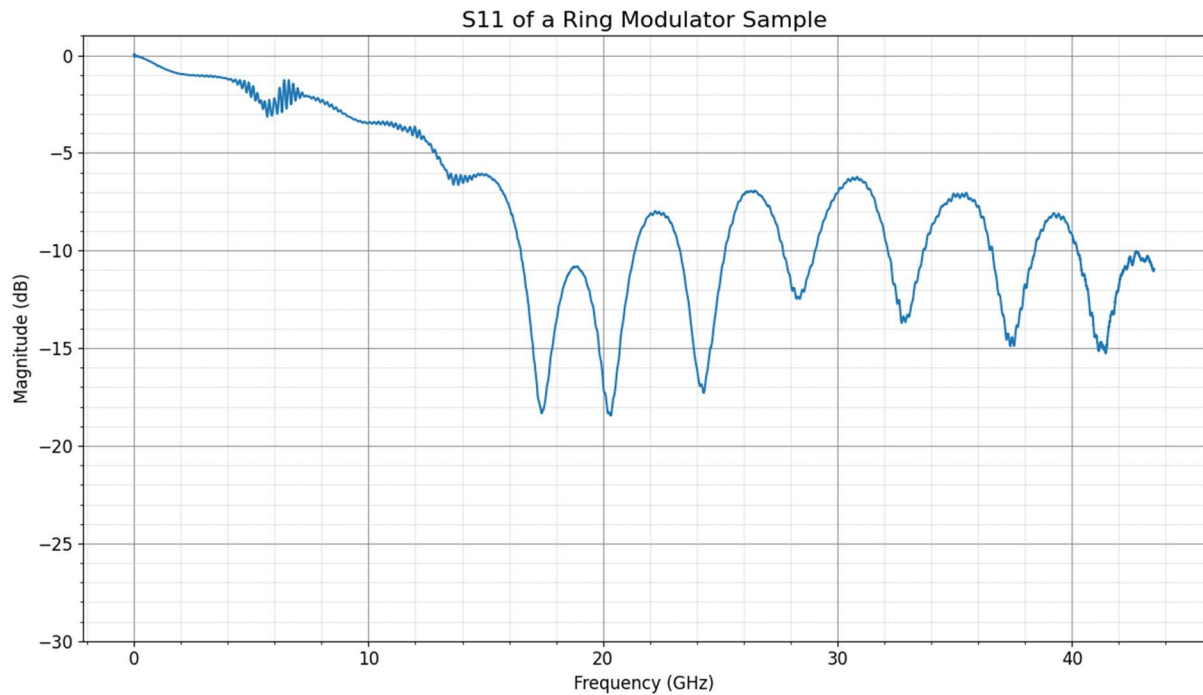


Figure 4.43: S_{11} of a ring modulator sample.

4.5.3 Mach-Zehnder modulator samples

The MZM sample builds differ from the RM builds mainly due to the different electrical requirements. Next to the fact, that two fan-out PCBs are needed, they also need to be equipped with electrical components. First the solder paste is dispensed on the component footprints with a pneumatic solder paste dispenser under a microscope. Then the components are placed onto the footprints. The soldering was done on a reflow station. The PCB before soldering can be seen on the right in figure 4.44. Here a GCPW design has been chosen. The components were equipped on the side of the PCB with the dedicated capacitor footprints for the RF return signal. The same configuration was chosen for all MZMs discussed in the next chapter.

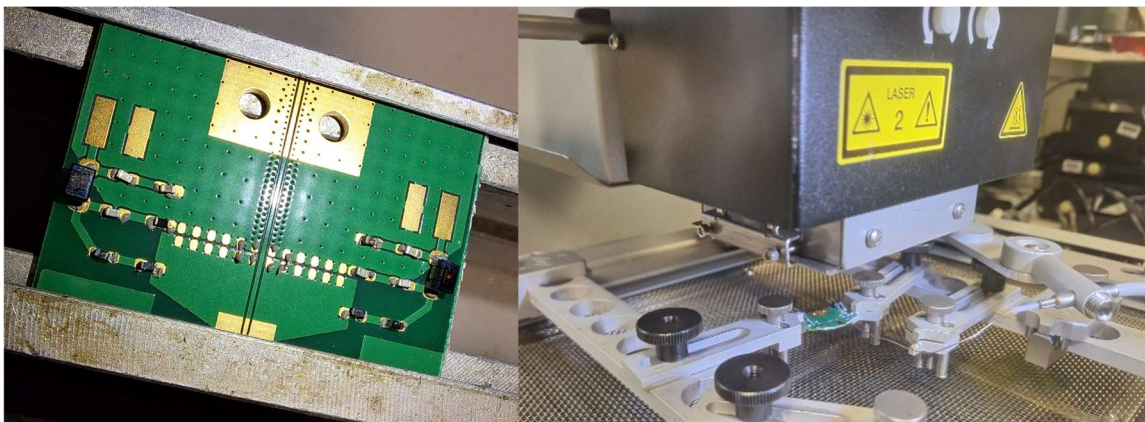


Figure 4.44: Left: An AN-2061 style fan-out PCB for Mach-Zehnder modulators before soldering. Solder paste for the smaller components has been dispensed and most components have been placed. Right: The same PCB fully populated inside the reflow station, used for soldering the components to the PCB.

These fan-out PCBs are then used in the sample building process as mentioned in section 4.5. As can be seen in figure 4.45, the COTTONTAIL chip has a non-ideal layout for wire bonding. The wire bonds to the termination fan-out PCB are extremely long. This is an intentional trade-off in the chip's layout, which will unfortunately decrease the RF response of the samples.

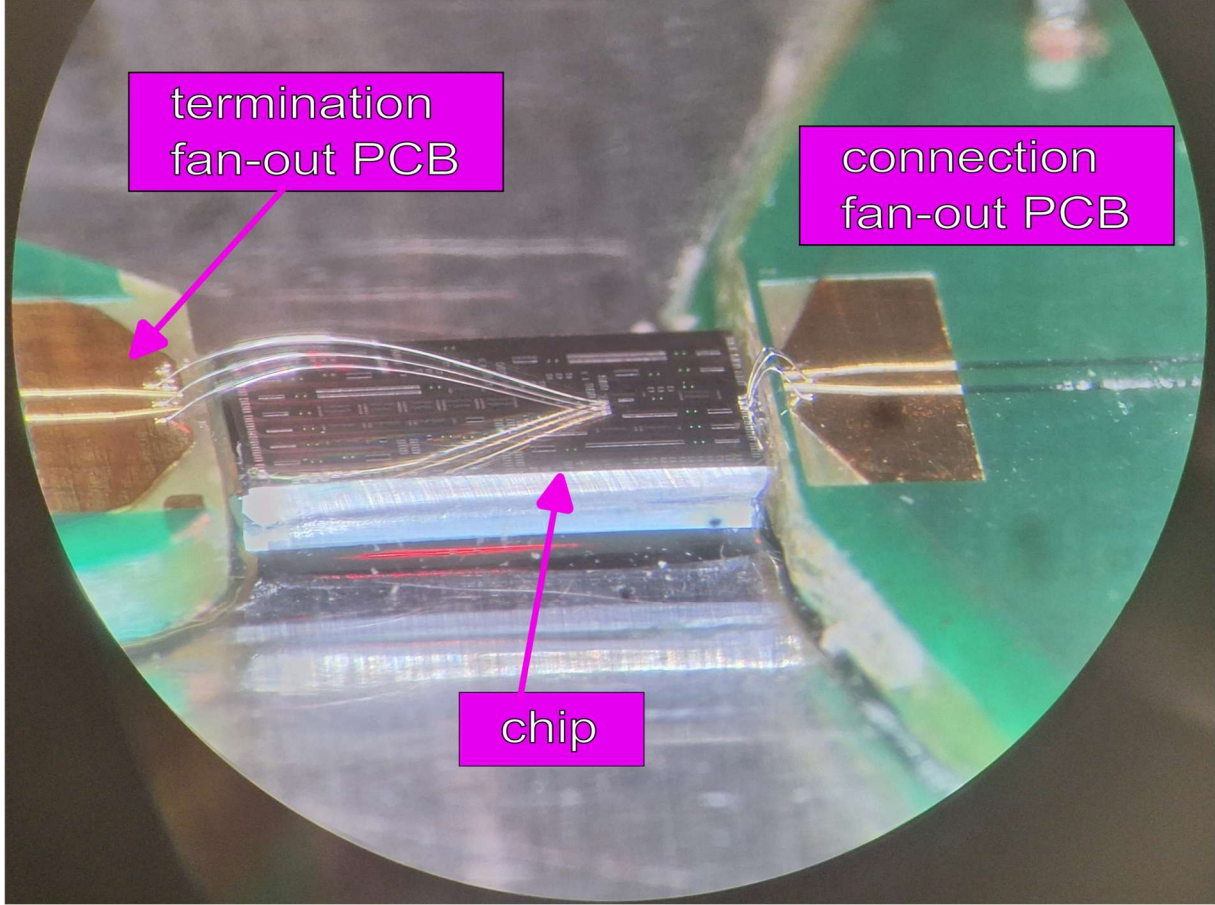


Figure 4.45: Mach-Zehnder sample build before the angle-polished fibers are attached. The wire bonds from the termination fan-out PCB to the chip need to be very long due to the chip layout.

Since the irradiation experiments started with RMs, no more usable glass-style fiber holders could be used. Therefore, another solution including the plastic resin printed holders had to be developed. It was decided to glue the fibers to the chip with the OP-54. If the fibers were to be completely parallel to the chip surface, this wouldn't mandate a change in the facet polishing angle. But an additional lateral offset Δx would occur compared to coupling without glue. To express this offset and the position x , where the beam center hits the chip, the terminology from figure 4.4 is kept. $x = 0$ is the position where the beam exits the fiber bottom. The resulting formulas are

$$x = \frac{d \cdot n_{\text{fiber}} \cdot \sin(\alpha)}{\sqrt{n_{\text{glue}}^2 - n_{\text{fiber}}^2 \cdot \sin^2(\alpha)}}. \quad (4.5)$$

Here d is the glue layer thickness. To show the difference $\Delta x = x_{\text{glue}} - x_{\text{air}}$ between air and glue (for air n_{glue} is changed to $n_{\text{air}} = 1$ in (4.5)), additional values are needed. For the fiber $n_{\text{fiber}} = 1.462$ is assumed (the same as the in section 4.2 mentioned Corning SMF28). For the cured OP-54 Dymax states

a refractive index of $n_{\text{glue}} = 1.506$. A realistic assumption for the distance between the fiber bottom and the projection plane is $d = 500 \mu\text{m}$. This results in $\Delta x = -34 \mu\text{m}$, which means the offset with glue is actually smaller than without it.

Due to the geometry of the MZM samples, the target overhang of 5 mm was exceeded considerably. Compounded by the adhesion force of the glue, this may lead to the fiber bending towards the chip surface. It remains to be investigated whether this effect is noticeable. The MZM samples were built at a time when no angle-optimized, angle-polished fibers were available. The fibers used were angle-polished on the modified chuck, whose angle did not undergo the optimization routine, unlike the full-plastic version shown in section 4.2.

The sample building process needs to accommodate the gluing of the fibers to the chip, next to the generally higher overhang requirements. The OP-54 is dispensed from a syringe. The exact model used doesn't allow for very small amounts of glue to be dispensed. Therefore, the glue was dispensed on top of the fiber, very far away from the GC. This leads to a small amount of glue creeping between the fiber and microscope slide towards the chip surface. If the correct amount of glue was dispensed, with some effort the fiber can still be finely positioned above the GC to maximize the coupling. Ideally the glue only spreads a bit further than the surface covered by the fiber. If too much glue was applied it creeps up the facet, avoiding TIR to occur therefore destroying any coupling. In this case the fiber needs to be removed, and the sample needs to be carefully flushed and cleaned with acetone. Since the wire bonds are already attached at this point, they may get easily bent or destroyed in this process. If coupling optimization was achieved, the glue needs to be hardened instantly with UV light. Often the curing leads to a temporary decrease in coupling efficiency. However, it will recover close to the original level, sometimes exceeding it. For the second side the process needs to be repeated. If too much glue was applied, the sample can still be cleaned at this stage. The result of this gluing step is shown on the left in figure 4.46.

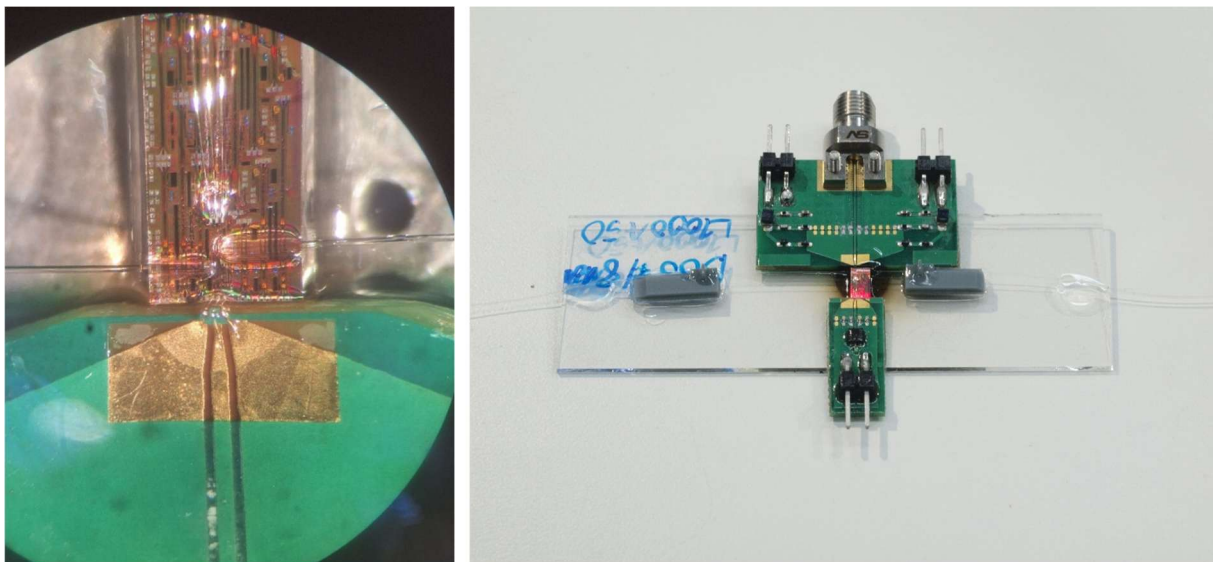


Figure 4.46: Left: Glued down fibers onto the top surface of the chip. Right: Complete Mach-Zehnder modulator sample. The sample was already irradiated, which results in the substrate discoloration around the chip.

The right of figure 4.46 shows a completed MZM sample. Summarizing the entire sample building process, it becomes apparent that many improvements to the packaging and interconnection technology must be made. Some of these difficulties couldn't be avoided since the COTTONTAIL layout is not optimized for sample builds with the testing MZM and RM structures. The chip's focus lies on versatility. For example, the long wire bonds on the MZMs and the increased overhang of the fibers come to mind. However, the presented solutions like glass-printed fiber holders and the gluing open possible paths forward. Since the intense X-ray irradiation measurements presented in the next chapter could be conducted, their viability was already proven.

5 Impact of X-ray irradiation on COTTONTAIL chip modulators

This thesis is the second one investigating X-ray irradiation damage to the COTTONTAIL chip and the first one doing so in-situ. The Bachelor thesis of M. Ludwig (Master cand.) [41] describes the first findings of the irradiation effects on the RMs on the chip. It was irradiated at a dose rate of $33 \text{ Gy}\cdot\text{s}^{-1}$. The measurements were conducted ex-situ. This means no samples like in 4.5.2 and 4.5.3 were built, instead a complete chip was irradiated and then characterized in the setup introduced in 4.1. It could be observed that the RMs exhibit a non-linear response with increasing TID. Up until 3 MGy no significant change in the EOE response could be observed. Then in the range up to 6 MGy the response dropped by 20 dB compared to the initial stable level. The wavelength that yielded the highest EOE response also shifted to higher wavelengths at higher TIDs, but only within a 1 nm range. Above the 6 – 7.5 MGy threshold, no further degradation was observed and all investigated RMs still were operational across the entire tested frequency range up to 40 GHz. Crucially the thesis also looked at forward bias (FWB) annealing. It was demonstrated that a current of $50 \mu\text{A}$ applied for a relatively short time of 10 s is sufficient to recover most of the modulation capabilities. The EOE response at lower modulation frequencies was observed to recover slightly quicker than higher ones.

5.1 Irradiation setup

For the in-situ measurements a setup like the one described in section 4.1 must be built around an irradiation chamber. The same irradiation chamber as in [41] was used. It is located at the Institute of Experimental Particle Physics (ETP) of the KIT. The radiation source is a 3 kW X-ray tube capable of generating up to 60 keV photons. The spectrum consists of a continuous part of Bremsstrahlung and a characteristic peak due to the ionization of the tungsten anode. In the experiments it is used with a current of 30 mA to maximize photon yield while still not inducing excessive wear. This results in a power of 1.8 kW.

Figure 5.1 shows the whole X-ray setup on the left. It shows the tube assembly at the top and a half automatic sample position stage at the bottom. The sample is placed on a lifting table on top of the automatic sample position stage. The latter will not be used in the measurements presented in this chapter. On the right an overlay on the X-ray tube is presented. Noteworthy is the distance to the sample, which is measured from the reference height and not the tip of the lead collimator pointing towards the sample. The Beryllium window seals the vacuum of the tube, while still letting most of the X-rays pass due to its low density, while the vanadium filter cuts away the lower energy part of the X-ray spectrum. Figure 5.2 shows the relative beam intensity distribution.

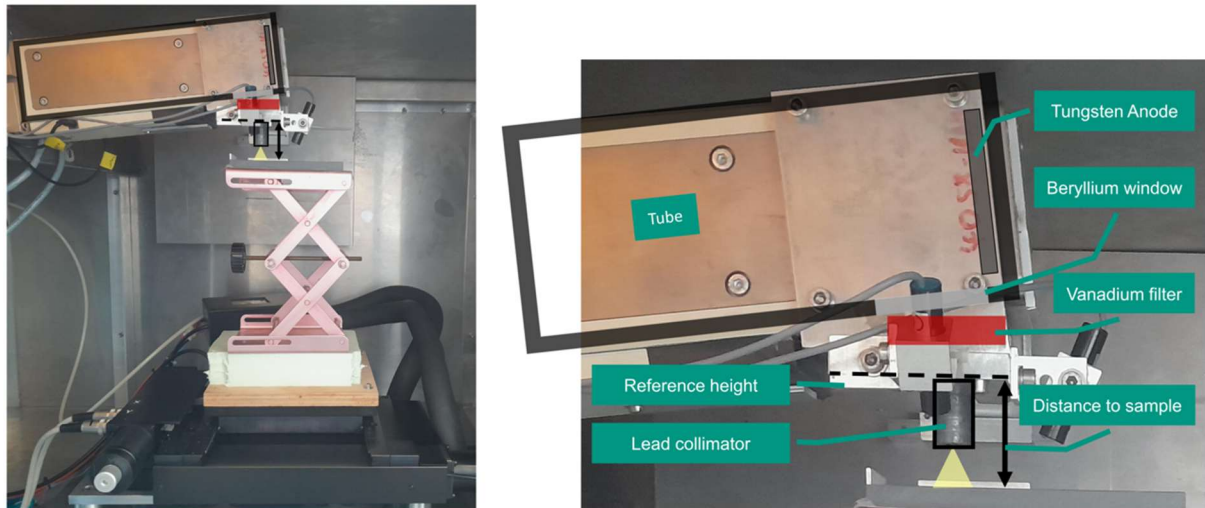


Figure 5.1: Left: Inside view of the irradiation chamber used for the irradiation experiments. The top part contains the X-ray tube. Underneath is the sample stage on top of a lifting table. Right: Close-up image of the tube and the sample stage (both taken from [41]).

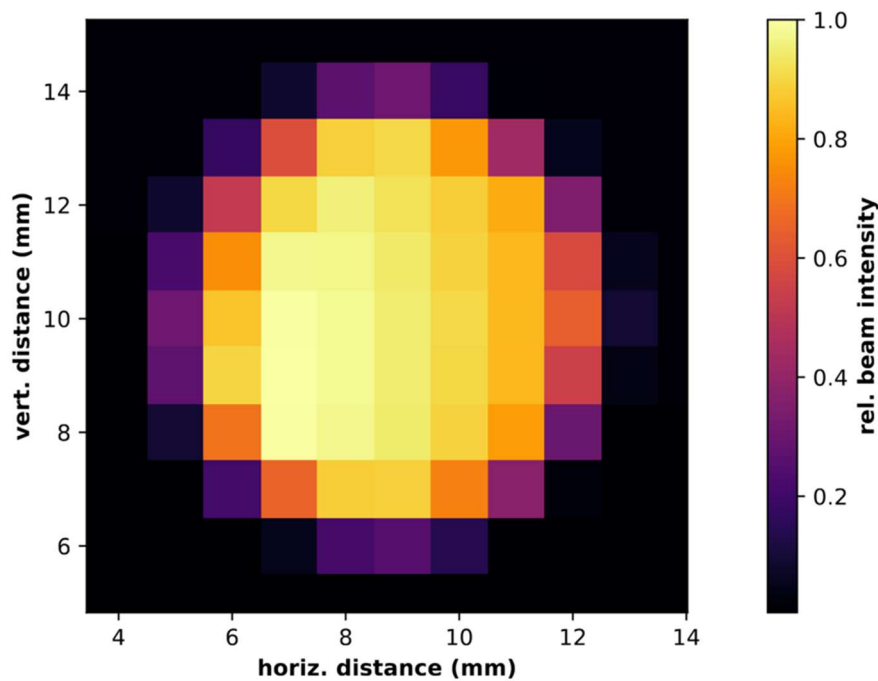


Figure 5.2: Relative beam intensity distribution under the collimator (taken from [41]).

Comparing the devices for the irradiation setup and the standard characterization setup introduced in 4.1 most devices are identical. The OPM was changed to an OPM-200-S-2 high-sensitivity optical power meter with two 2 mm InGaAs detectors (all Santec Holdings Corp.). The power supply for biasing and annealing is technically an identical device but was changed to a B2912B precision SMU from Keysight Technologies. Since the DC (wavelength sweep) and RF (frequency sweep) measurements are time intensive they are automated. To achieve this functionality, a FFSW 1x2 fiber optic switch (Agiltron Inc.) is used, which either leads the modulated light to the fast photodiode connected to the SA or the OPM. The device interconnections are done via LAN with a network switch and USB for the optic switch. A laptop, which controls the setup and records all data, is also connected to this switch. Figure 5.3 shows all the optical, electrical and LAN connections.

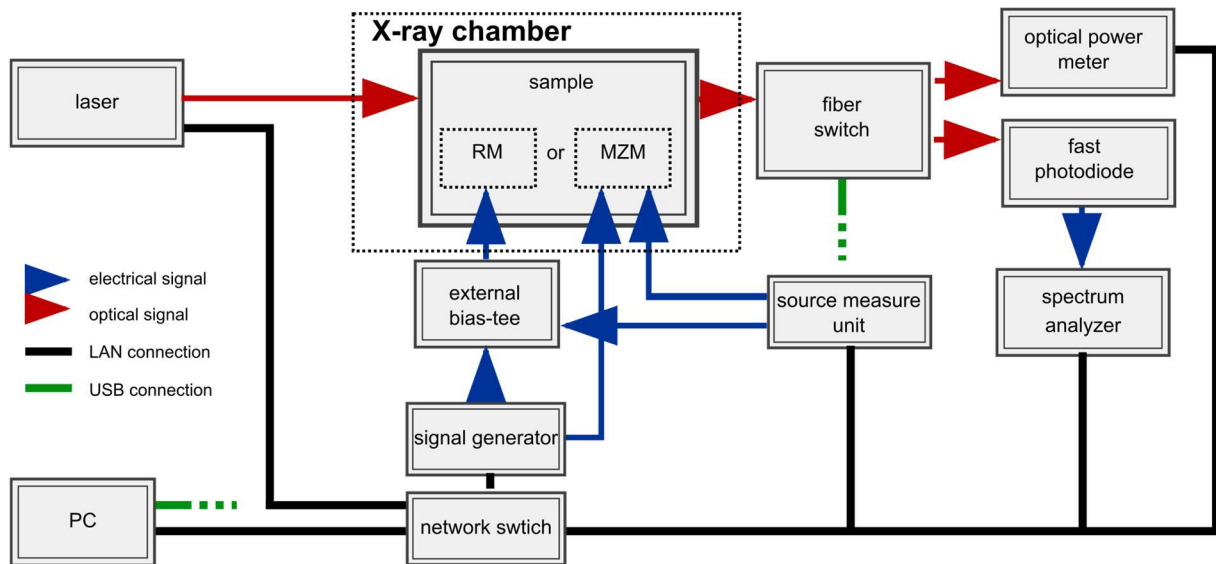


Figure 5.3: Schematic of the irradiation setup. The reference clock connection between the signal generator and the spectrum analyzer was omitted here.

For the setup a new Python code base with device drivers, real-time data plotting and data acquisition was created. The DC and RF sweeps follow the same program logic as shown in figure 4.5 from section 4.1. No active control for the previously observed working point shift (which is most likely due to either radiation or temperature) is present. Therefore, DC sweeps need to be repeated periodically, and the new optimal working point needs to be determined. Otherwise, no distinction between a working point shift or modulation capability change can be made. The repetition criteria chosen for all presented measurements was a drop in the EOE response by 3 dB at a reference electrical RF frequency of 100 MHz. One RF sweep here means that the EOE response of a predetermined list of frequencies is measured once.

The automatic optimal working point algorithm follows the flowchart shown in figure 5.4. After a DC sweep the laser wavelength is set to an expected value of the working point. The RF reference frequency is applied and the resulting EOE response is the initial maximum value. For RM samples from the detected notch position closest to 1550 nm a fixed offset of 0.6 nm is applied. This value was found to result in the most reliable behavior. Therefore, the modulation will take place on a rising flank of a notch. In the case of MZMs, from the closest peak in the DC spectrum to 1550 nm an offset of one fourth of the distance to the next peak is applied. This positions the working point on a falling flank close to the point with the highest modulation sensitivity. In both cases the working point then needs further optimization. This is done through an iterative process with an increasingly smaller step size (0.25 nm, 0.1 nm, 0.01 nm and 0.001 nm). From the wavelength where the previous maximum was, it is checked if a step upwards increases the EOE response. If that is the case, it will be increased until the response gets smaller. If an increase was not detected at the first step to higher wavelengths, a step down is made. If this step down then yields an increase, it is repeated until the EOE response decreases again. If a new preliminary optimum was found or the initial one before the current step size was already the highest detected value in both wavelength directions, the process is started anew with the next smaller step size.

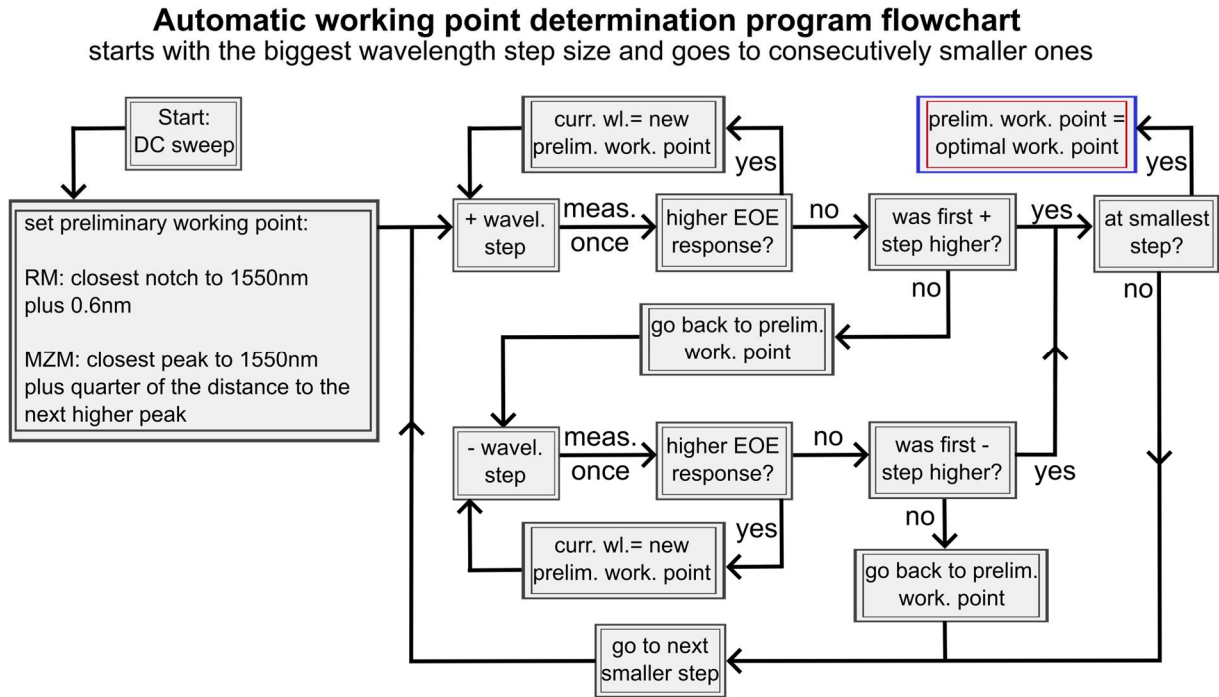


Figure 5.4: Flowchart of the automated method to find the optimal working point for a given modulator type.

This optimization is part of the larger program flowchart, which is shown in figure 5.5. It consists of two parts: a pre-irradiation measurement and a continuous measurement under active irradiation. In the pre-irradiation measurement a DC sweep is taken, the optimal working point is determined and one RF sweep over all modulation frequencies is recorded. When the irradiation chamber is ready, the continuous measurement is started via the GUI. After a timer has passed, the shutter of the X-ray tube is opened manually. Directly a new DC sweep will commence, followed by the optimization step and RF sweeps. Repeating the DC sweep immediately was chosen to compensate for potential temperature-induced changes due to the irradiation. As previously mentioned, the RF sweeps will now continue until the threshold has been reached. If that is the case, the process starts again with the next DC sweep.

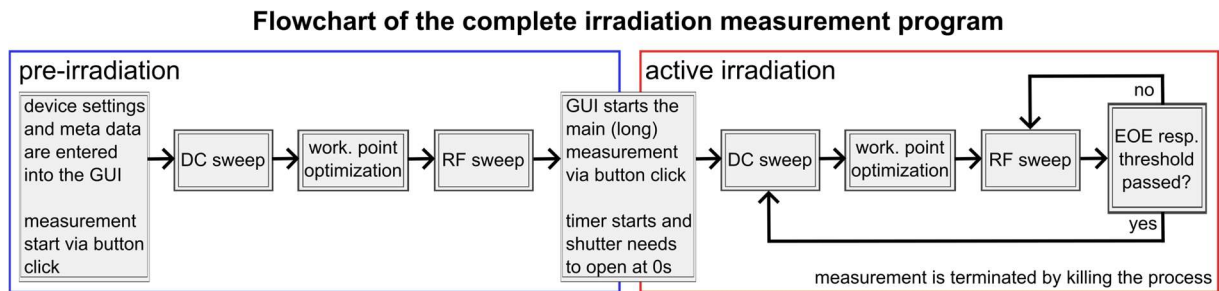


Figure 5.5: Flowchart of the complete irradiation measurement program.

In total the program uses six driver classes, a main.py (due to time limitations MZMs and RMs use two slightly different versions), which starts a live plotter and measurement routines class (the latter also has two slightly different MZM and RM versions). The latter also handles data acquisition, where the data is written directly to a csv file, which is then saved. Due to simplicity the live plotter only accesses the already saved data, therefore a small lag between recording and displaying exists. The measurement routines class contains higher level device procedures and the program flow logic. Both, the live plot

and measurement routines are started in different threads from the main.py, which also handles the GUI. Figure 5.6 shows the measurement GUI (RM version). Figure 5.7 shows an example of the live plotting.

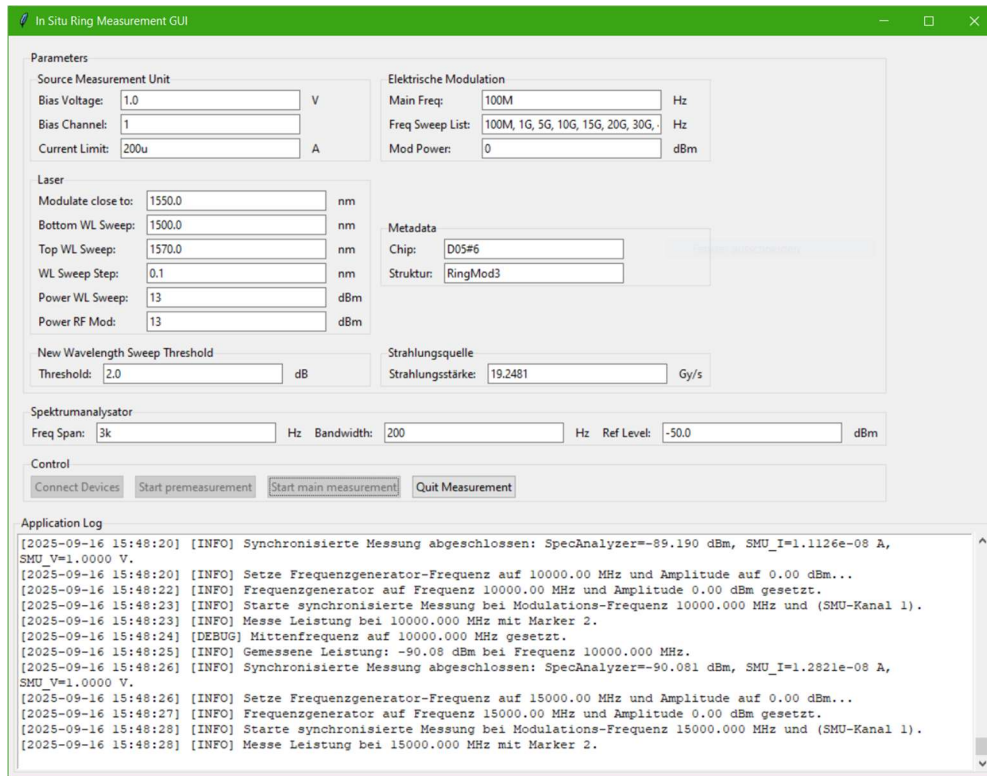


Figure 5.6: The GUI of the in-situ measurement software during a measurement.

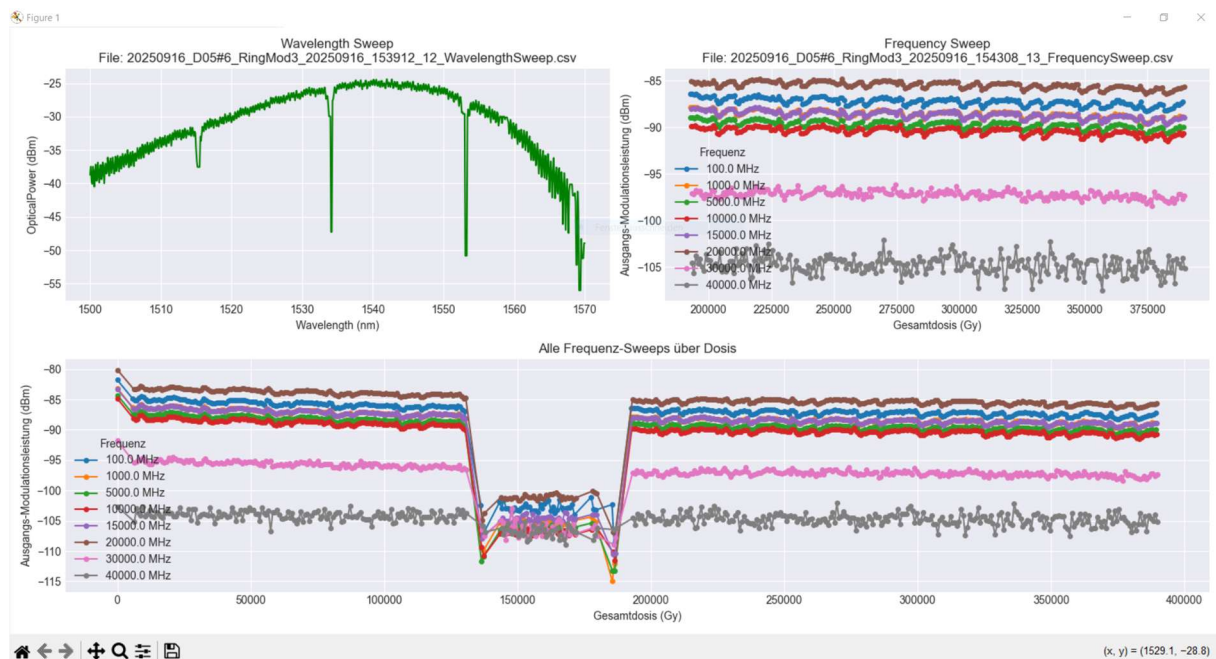


Figure 5.7: The live plotter of the in-situ measurement software during a measurement.

With this preparational work the setup could be integrated into the irradiation chamber environment. This is shown in the following figures.

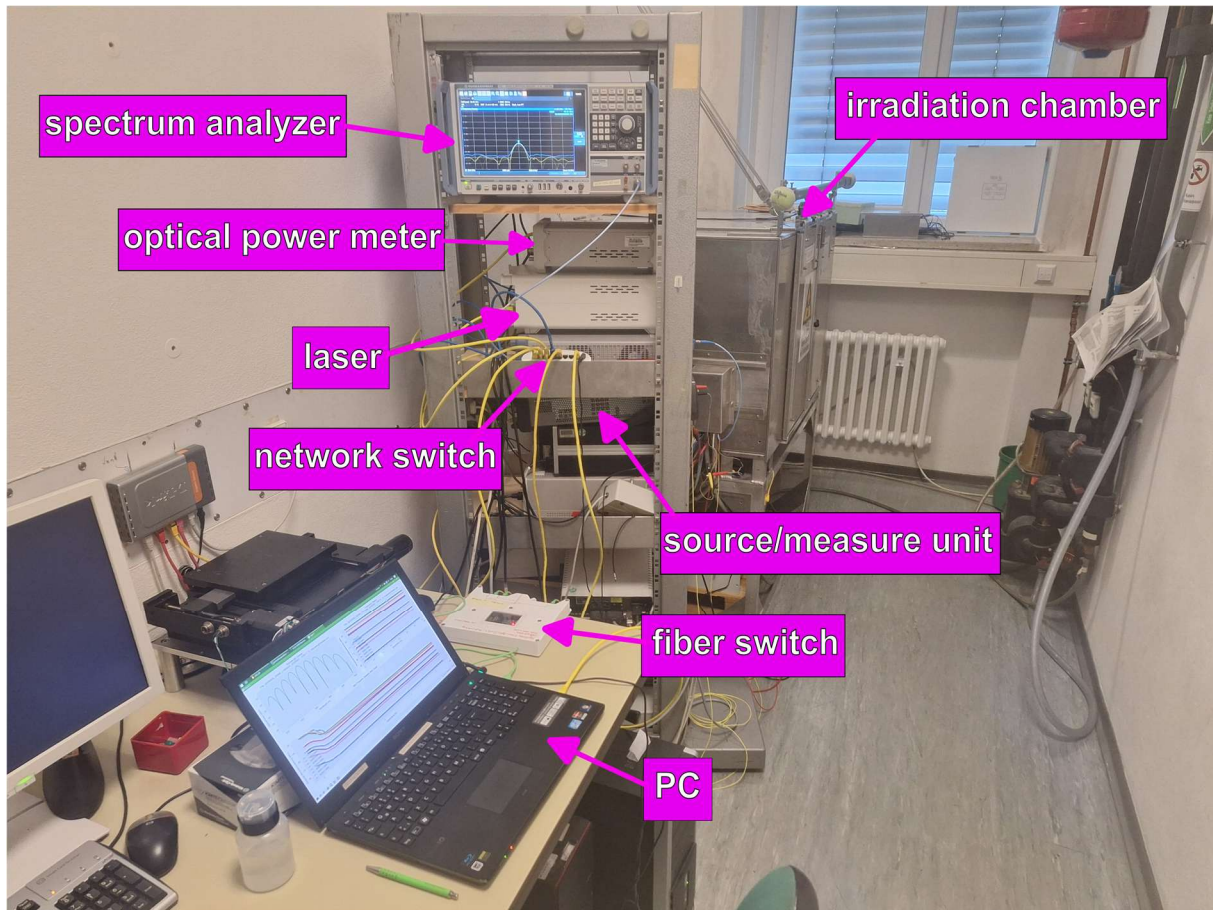


Figure 5.8: The complete irradiation setup during an active measurement.

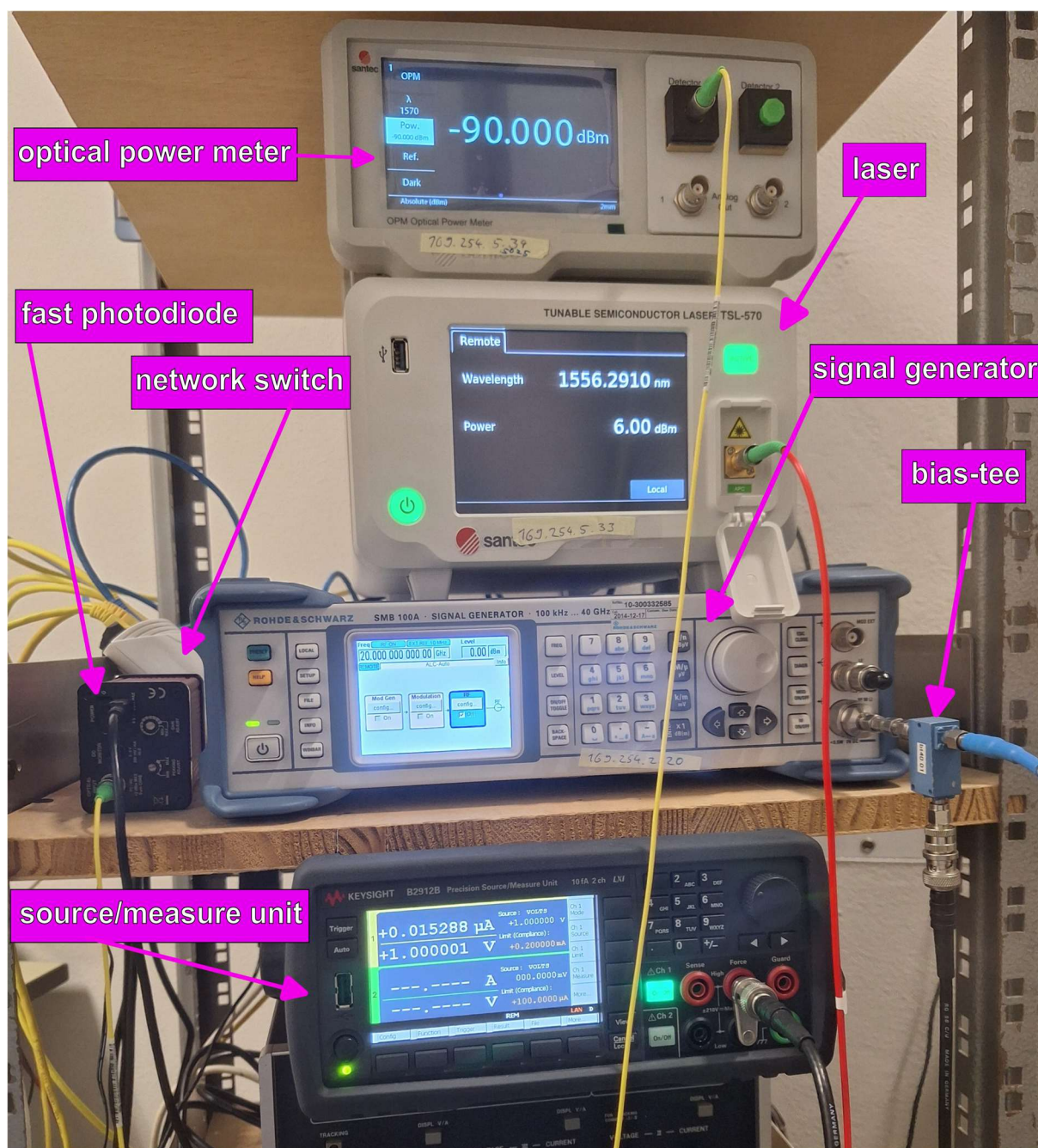


Figure 5.9: All the measurement devices (except for the spectrum analyzer) during an active measurement.

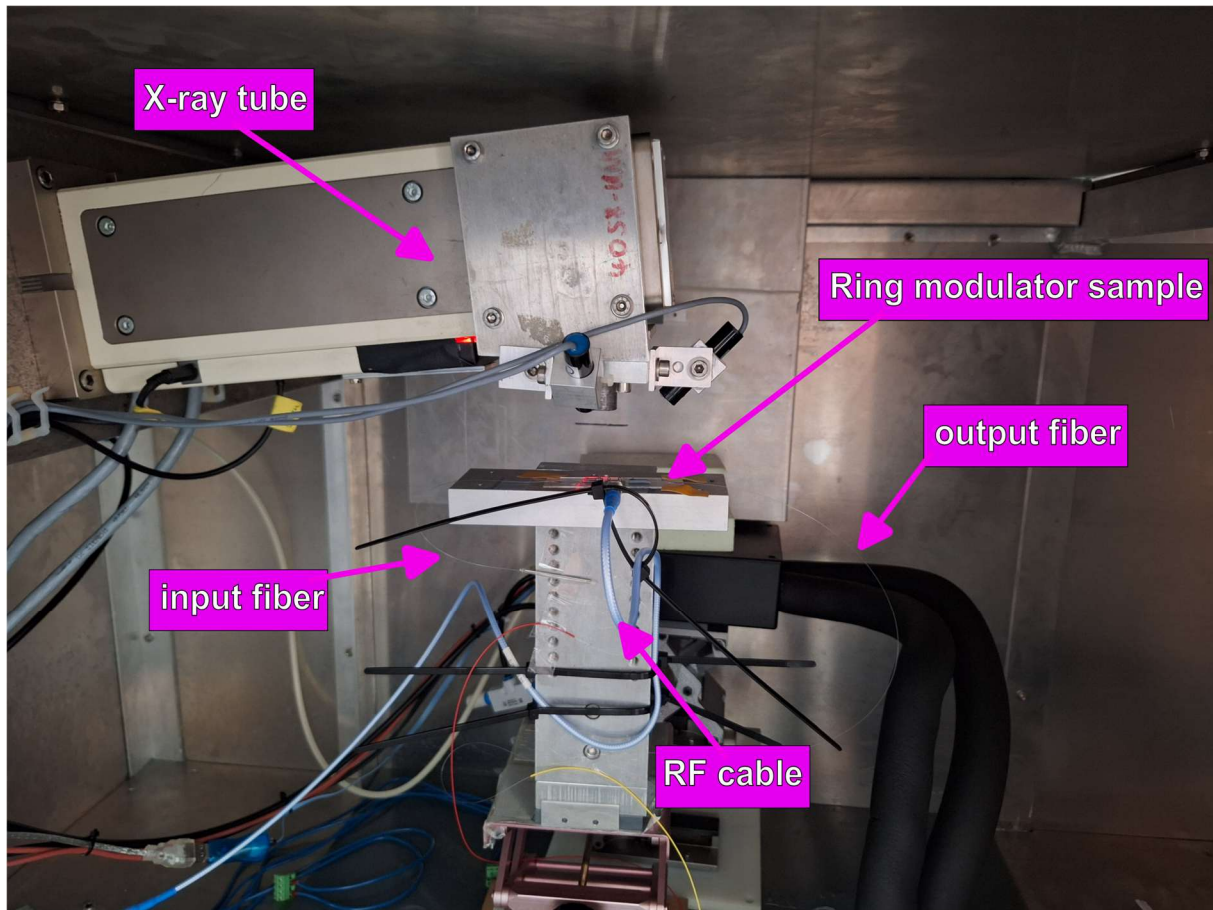


Figure 5.10: A ring modulator sample inside the irradiation chamber just before measurement start.

The program largely performed as intended, however some changes and bug fixes are in order. First, an option to skip the pre-irradiation measurement should be introduced, since sometimes it is more convenient to directly continue with the irradiation. As it will become apparent in 5.3 and 5.4 that the working point shift is not as large as foreseen. Therefore, the first DC sweep after a pre-measurement may be skipped. Also, to resolve the drop of modulation due to the TID without interruptions, which come from passing the simple threshold, a “smart” criterion for DC sweep timing might improve the datasets. This could be done for example with a minimum time between them. Also, it should be asked if a full DC sweep is always necessary. Additionally, the MZM and RM specific code should be unified. This step wouldn’t need a lot of changes, beyond a toggle switch in the GUI. Finally, the stop button for the measurement doesn’t work as intended. Most likely due to a problem with the thread stop-call handling within the called thread (measurement routines). Currently, the whole process must be killed, which is not ideal for the devices and data acquisition.

5.2 Fiber degradation

The first irradiation measurement presented does not yet look at the RM or MZM behaviors, but the one of the optical fibers used. For all sample builds standard Germanium-doped PM and non-PM SMFs were used. The concern is that typically for irradiation environments specialized radiation-hardened fibers are utilized. These are pure silica core fibers with Fluorine-doped cladding. It must be investigated whether radiation-induced attenuation (RIA) in the used Germanium-doped fibers influences the measurement results. RIA occurs due to point damage creating color centers [49]. The effect can be observed on the

microscope slides used for the sample builds, like shown in figure 5.11. The picture is taken from a disassembled sample's microscope slide, where the radiation shadows of the GCPW RM fan-out PCB and the fiber holders are clearly visible. Also, through a microscope the shadows of the fibers themselves are slightly visible.



Figure 5.11: Radiation-induced discoloration of a microscope slide of a disassembled ring modulator sample due to color centers.

A simple sample was built where a PM and a non-PM fiber were glued directly next to each other onto a microscope slide. Prior to gluing, the coating and buffer layers were stripped so that only the core and cladding were irradiated, analogous to the standard samples (regarding primary irradiation). The resulting sample is shown in its pre-irradiated and irradiated state in figure 5.12.

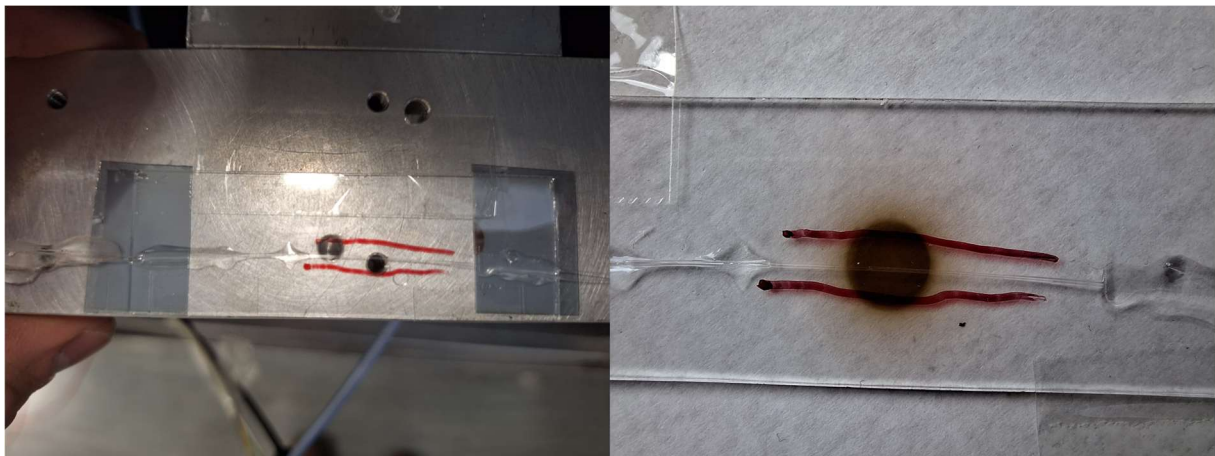


Figure 5.12: Left: Sample consisting of two fibers on a microscope slide before irradiation inside the irradiation chamber. Right: The same sample after irradiation.

For this measurement only the laser, fiber switch and OPM are needed, where for the latter both detector channels are used. A simple program was created to take a DC sweep measurement of each fiber consecutively every five minutes, utilizing existing drivers from the standard irradiation measurement

code base. The laser was set to 6 dBm since the setup itself has some losses and the other irradiation measurements used the same setting. The spectrum is recorded in 0.1 nm steps. The first spectrum of each fiber is recorded without irradiation. The sample is positioned 25 mm away from the reference height.

The results of these measurements are plotted in figure 5.13 (PM SMF) and in figure 5.14 (standard SMF). The measured TID goes up to 10.92 MGy in two days, 14 hours and 35 minutes measured. The dose rate was estimated with $48.463 \text{ Gy}\cdot\text{s}^{-1}$, based on a calculation tool on a computer connected to the irradiation chamber. A total of 255 spectra for each fiber type were taken. In the figures low TIDs are indicated with blue colors, which turn red with increasing TIDs. Both distributions are not perfectly flat but are contained within a 0.19 dBm window. The PM SMF graph has some sudden spikes to higher and lower levels, while the standard SMF graph lacks these. The optical power in the PM SMF plot has a resolution of 0.001 dBm while the other has 0.01 dBm. Some data points far lower than -20 dBm were ignored in the dataset (only -90 dBm was observed, but the code threshold for the plotting was set to -20 dBm).

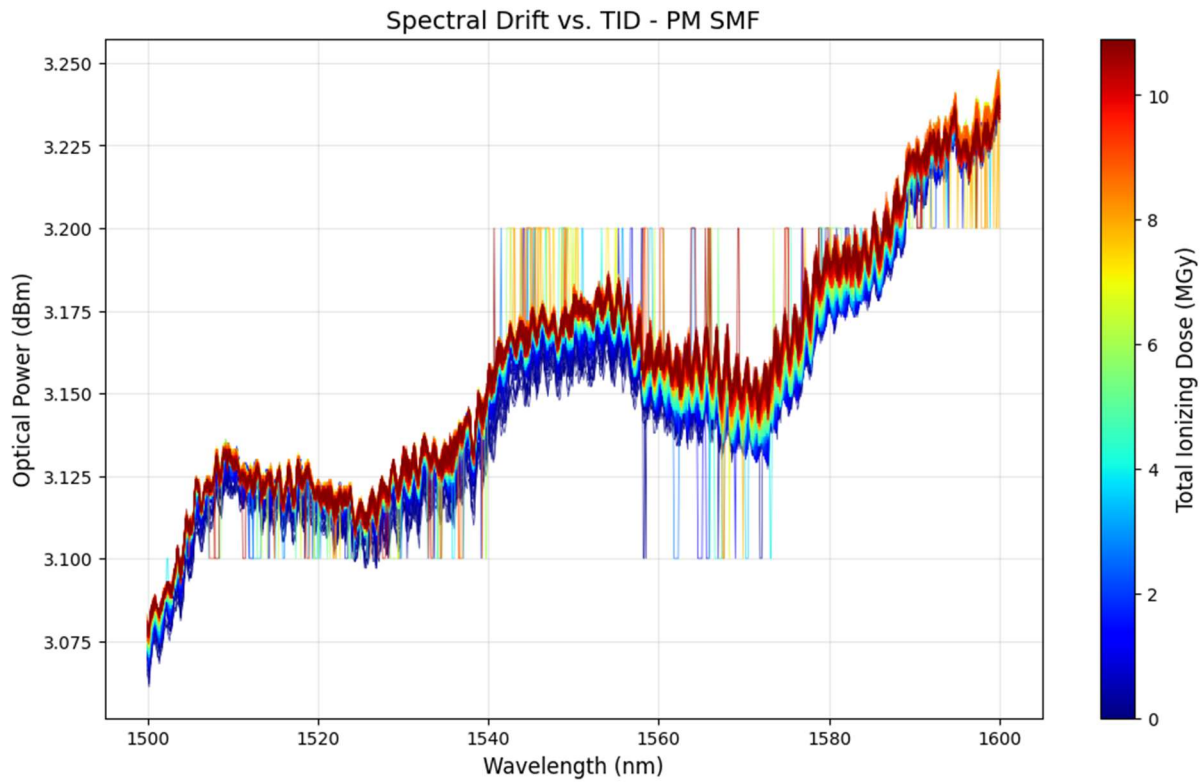


Figure 5.13: Spectral drift of a polarization-maintaining single-mode fiber at increasing total ionizing doses. The spikes occur due to the auto-range function of the optical power meter.

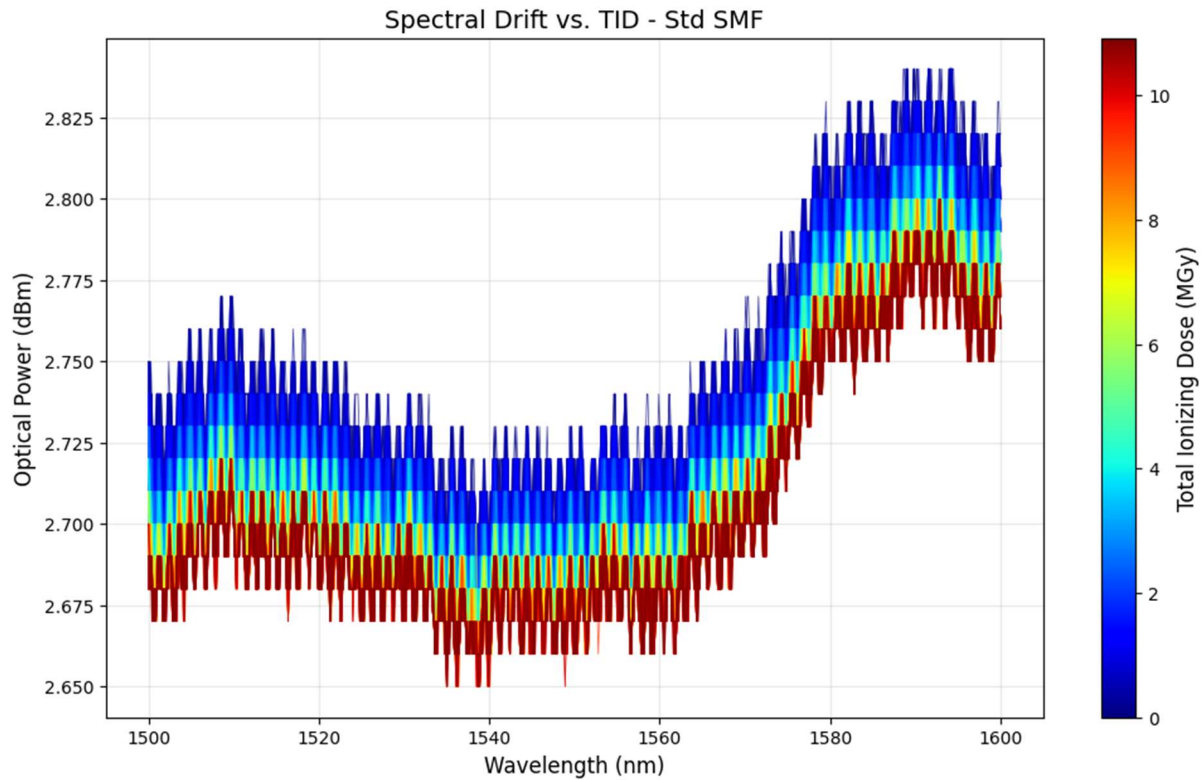


Figure 5.14: Spectral drift of a standard single-mode fiber at increasing total ionizing doses. The resolution is lower compared to the results in figure 5.13 due to the auto-range function of the optical power meter.

From an ideal fiber flat spectra would be expected, under the condition that the laser generates such a spectrum. The general shape of the spectra, which doesn't change significantly with the TID, is due to an ensemble of factors. These most likely are primarily non-ideal fibers and measurement equipment induced noise and calibration mismatches. The optical power resolution difference is an unintended consequence to setting the OPM auto-range functionality to "True". It appears that 3 dBm is the threshold value for the OPM to decide between two different amplification circuits. This explains, next to the two different measurement resolutions, the omitted -20 dBm data points, since the OPM couldn't switch between the circuits fast enough. Also, the spikes in the PM SMF curves can be explained by this. Clearly, there are jumps to 0.1 dBm discrete steps. The operation close to the 3 dBm threshold makes this theory even more probable. The color gradient demonstrates the expected decrease in optical power with rising TIDs for the standard SMF fiber. However, the opposite can be observed with the PM SMF. This is an unexpected result. A closer investigation of this behavior for selected wavelengths (1500 nm, 1550 nm and 1600 nm) is shown in the plots in figures 5.15 and 5.16. Here the relative change in the optical power is plotted over TID. As can be seen the RIA is contained to -0.05 dB for the standard fiber while PM SMF optical power values do in fact increase or at least remain stagnant. Technically, RIA would be plotted with inverted signs, but the logic from the color gradient of the previous figures aligns with the convention chosen here. It is expected that the conflicting RIA trends of both fibers are most likely caused by temperature sensitivity within the two amplification circuits (and their calibration curves) of the OPM. A further observation, which likely did not influence the experimental results, concerns optical crosstalk. When the laser was active in one fiber, the OPM detected power levels of approximately -85 dBm on the channel of the adjacent, non-active fiber. This

is most likely caused by leakage within the fiber switch into the inactive channel. When the laser is not active the OPM detects its minimum value of -90 dBm.

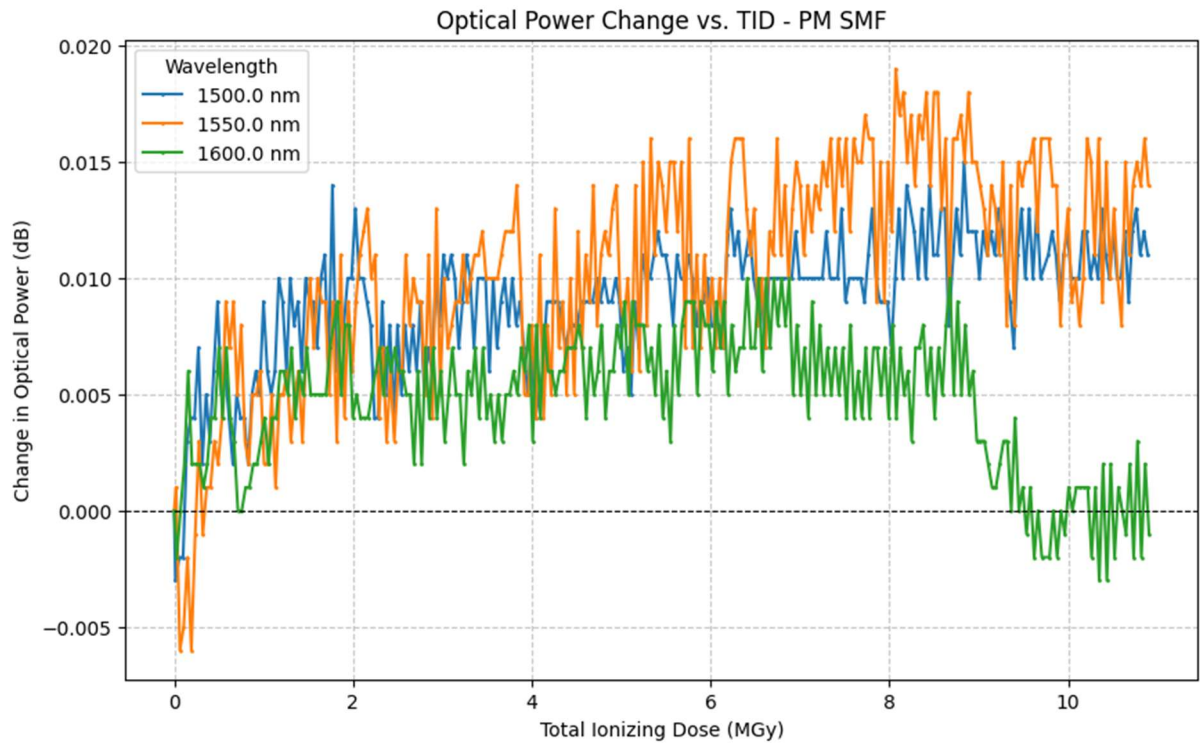


Figure 5.15: Change of the received optical power by the optical power meter through a polarization-maintaining single-mode fiber at increasing total ionizing doses.

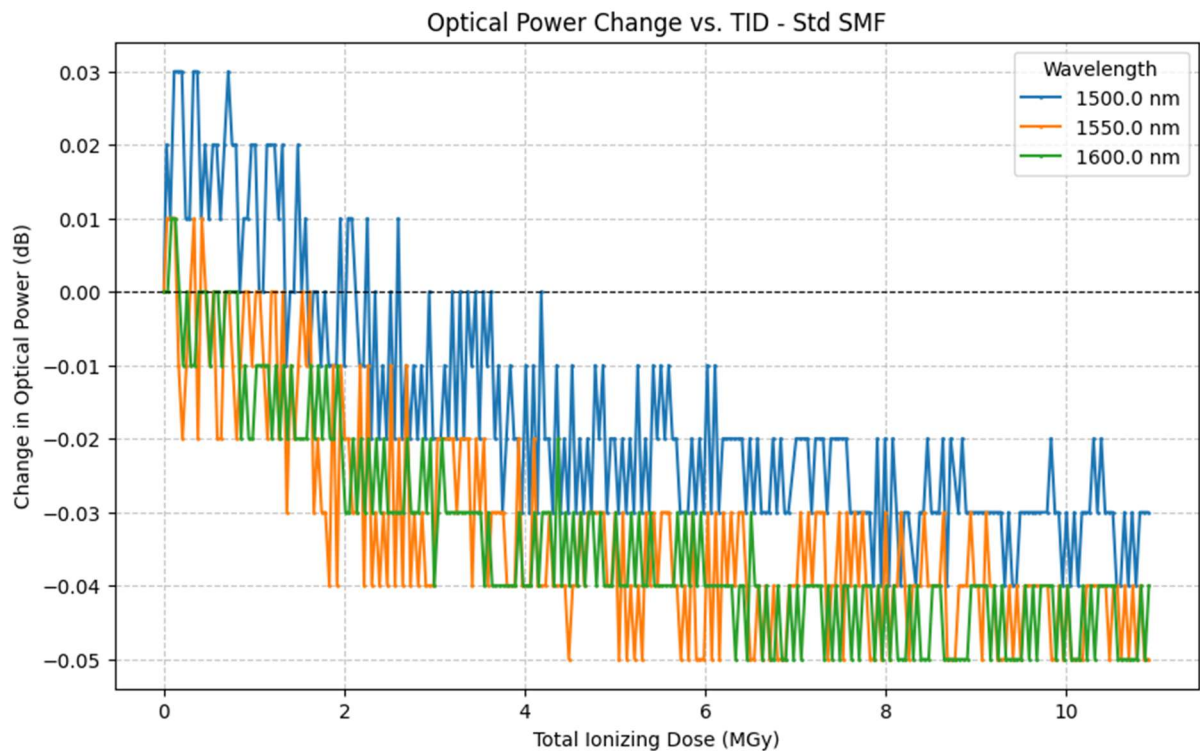


Figure 5.16: Change of the received optical power by the optical power meter through a standard single-mode fiber at increasing total ionizing doses.

According to these results RIA is not a factor with these fibers for the expected/given doses. With the clear limitation of the primary beam spot being very small, i.e. a diameter of approximately 7.7 mm. Therefore, for the irradiation experiments with the samples RIA doesn't need to be considered. This doesn't mean that for a final application in a detector environment, these fibers wouldn't suffer from RIA. There they would be irradiated with primary X-ray irradiation of much greater fiber lengths. But if the flat coupling would be pursued and it would become apparent, that Fluorine-doped fibers cannot be angle-polished or would be disqualified due to any reasons, standard fibers can be used for the coupling region up to a maximum length. The latter is to be determined by the maximum acceptable RIA.

5.3 Ring modulators

The first irradiation measurement was performed on the device DO5#9#RingMod4 (Wafer#Chip#Structure) with a GCPW RM fan-out board. The measurement lasted approximately six days, 20 hours, and 30 minutes in total, reaching an accumulated TID of 25 MGy. The dose rate at 25 mm is $48.5 \text{ Gy} \cdot \text{s}^{-1} \cdot 0.9 = 43.7 \text{ Gy} \cdot \text{s}^{-1}$. The factor of 0.9 compensates for the positioning within the beam spot, as shown in figure 5.2 in section 5.1. The laser was set to 6 dBm. The SA's bandwidth was configured to 300 Hz, the span to 3 kHz and the reference level to -40 dBm . The RM was reverse biased at 1 V. These values result from a manually conducted pre-measurement without irradiation. Figure 5.17 shows all the recorded frequency data and figure 5.18 shows all DC sweep data.

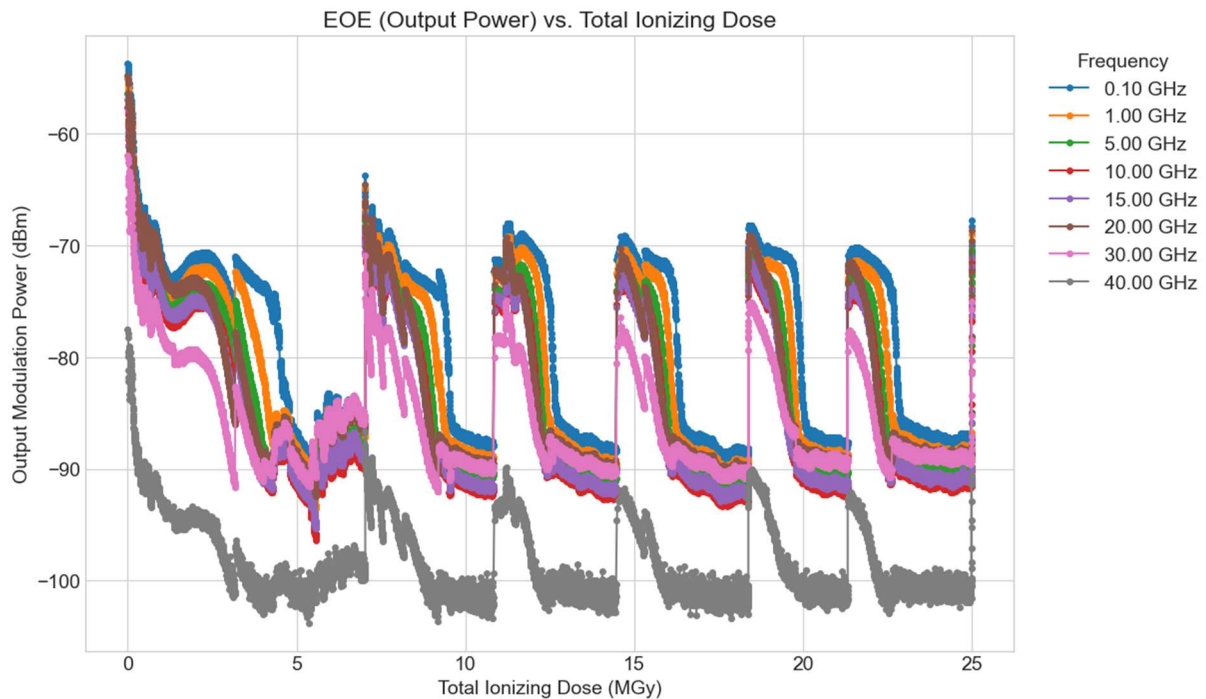


Figure 5.17: Electrical-optical-electrical response of the ring modulator sample for different modulation frequencies and for an increasing total ionizing dose.

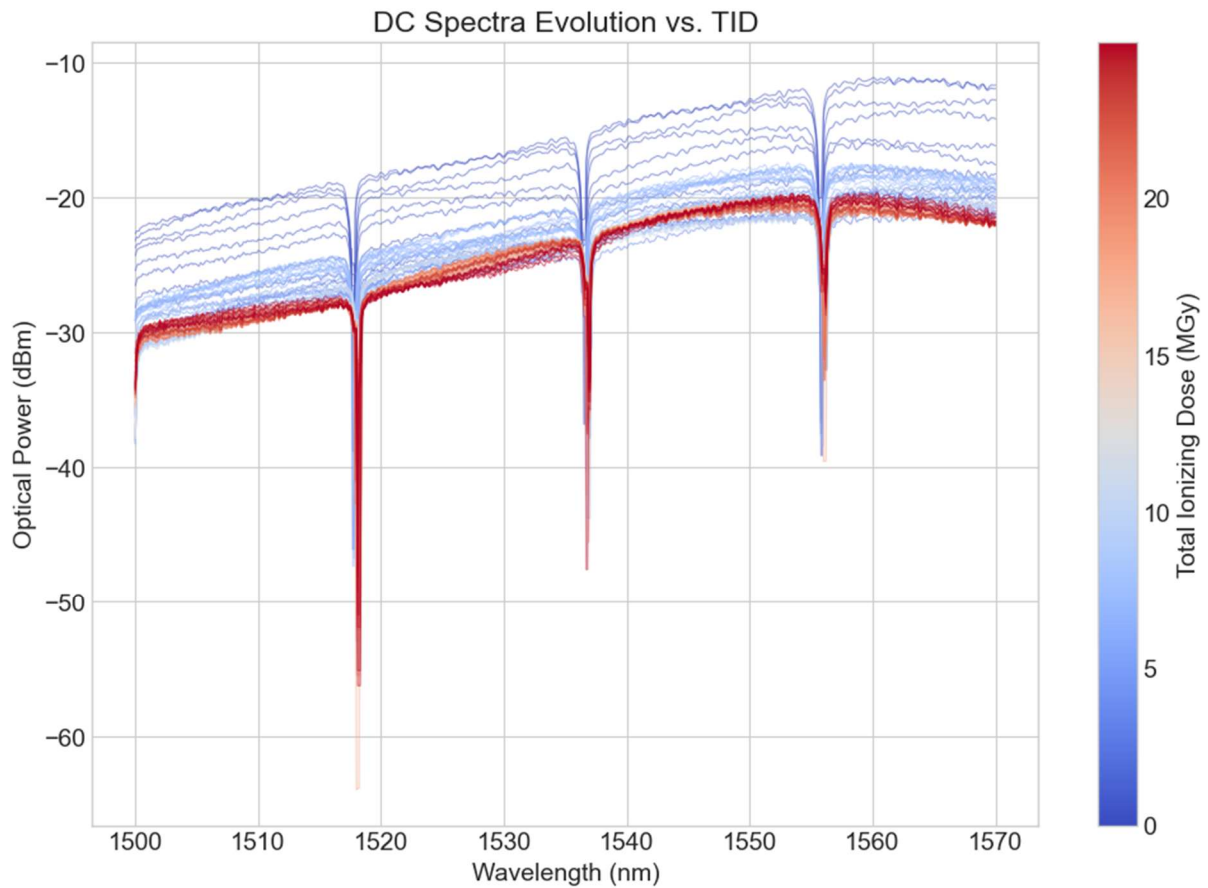


Figure 5.18: DC spectra evolution of the ring modulator sample for an increasing total ionizing dose. A version with inverted plotting order is shown in figure 7.5.

The RF data initially shows a peak EOE response of roughly -54 dBm for 100 MHz, while higher frequencies have an increasingly lower starting point. One exception is the 20 GHz curve, which experiences a jump to a higher initial EOE response compared to its neighboring frequencies. All frequencies' EOE response then deteriorates to the -70 dBm range, where it generally stays throughout the measurement, if irradiation damage does not dominate. The level of the modulation collapse is at around -90 dBm for all frequencies except for 40 GHz. The latter collapses to below -100 dBm, which is around the -110 dBm noise floor of the SA, leading to a noisy EOE response. It was annealed six times with a forward bias (FWB) at currents ranging from $50\text{ }\mu\text{A}$ to $200\text{ }\mu\text{A}$. The timestamps (or TID levels) corresponding to the annealing processes are identifiable where the EOE response recovers from the collapse level back to the -70 dBm range. Initially, all curves exhibit fluctuations ("swings") in their EOE response, which decrease over time. To compensate for the effects of the two RF cables used, calibration measurements were performed by M. Schneider (PhD). The measurement results are shown in figure 7.4 in the Additional measurement data section. The calibration curve was derived from measurements in a setup with three cables; therefore, the compensation for two cables (and one cable) was derived mathematically. The value for the respective frequency is simply added to the EOE response. For all following RF analyses this corrected dataset was used.

The spectral range of the DC sweeps was limited to a 1500 nm to 1570 nm range. This lessens the time requirements for each sweep. In total 95 curves were recorded. In figure 5.18 curves taken at lower TIDs are blue, whereas red indicates higher ones. The spectra show three clear peaks at roughly 1518 nm,

1537 nm and 1556 nm. Some curves, especially at higher TIDs, indicate a peak close to 1500 nm. Similar to the RF measurements, an initial drop could be observed from approximately -11 dBm at 1560 nm to levels around -20 dBm. This 9 dB lower power at the OPM alone transforms to a 18 dB lower RF signal measured with the SA at the RF measurements.

To analyze this data first it must be discussed how to deal with the swings in the EOE response plot and what is chosen as reference point for further measurements. The DC sweep plot in figure 5.18 shows clearly that the optical coupling efficiency in the beginning is initially radically deteriorating. Figure 5.19 shows this behavior for fixed wavelengths over the TID.

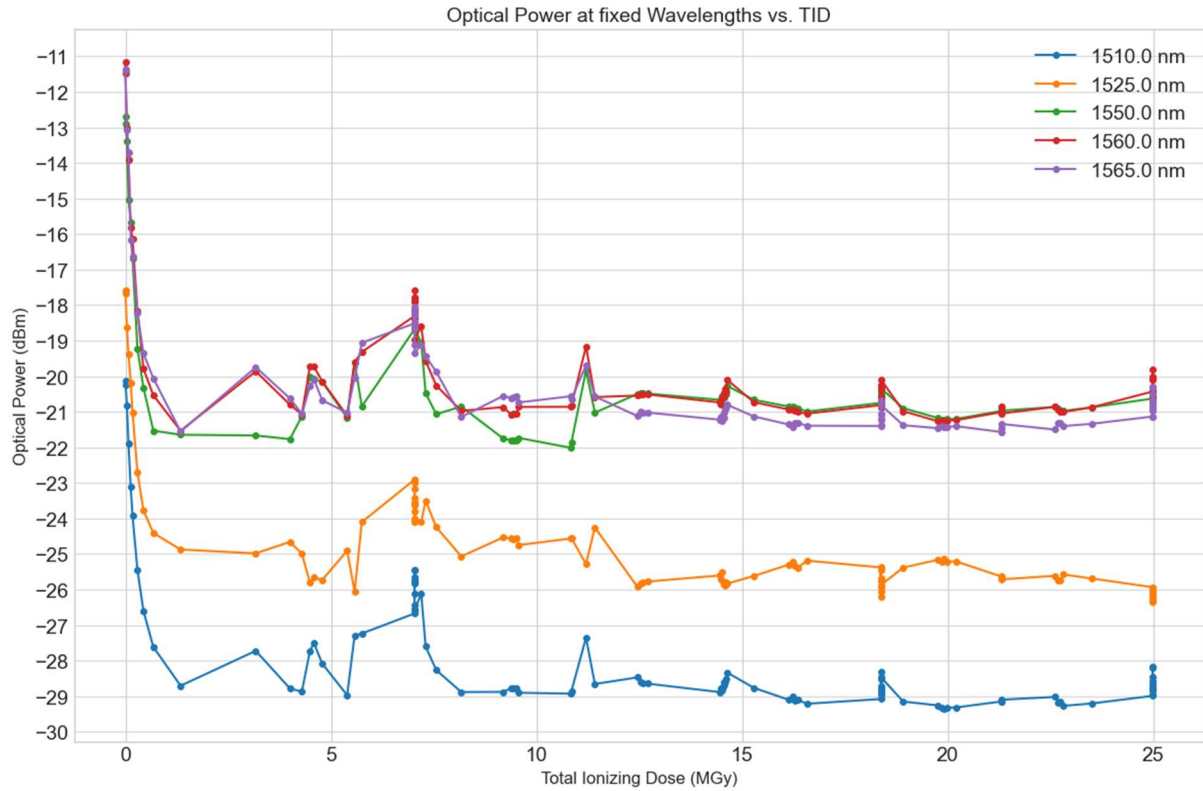


Figure 5.19: Coupling efficiency evolution of the ring modulator sample for increasing total ionizing doses.

This is caused by the glue still curing through the irradiation-induced heat. Optimally, the first data point while the sample has never experienced irradiation would be taken as a reference. Then several aspects like how the sample recovers its original modulation capabilities or where it crosses certain attenuation thresholds could be made. However, the initial dip combined with the subsequent noise prohibits this approach. Since the modulation capabilities are not expected to decline drastically, the most stable campaign was chosen as the reference. This means all following discussions assume the respective peak EOE response (single data point, not from the fitted curves) at each frequency within the range 20 MGy to 24 MGy to represent full modulation ability. This most likely shifts values but will not fundamentally change the findings derived from them.

Going from these reference points it can be shown, that the 100 MHz curve stayed over two attenuation threshold values of 3 dB and 6 dB for TIDs larger than 11 MGy. All results are shown in table 5.1. These points are defined by the last point of the filtered frequency curves (Savitzky-Golay: window length of

45 data points and a second order polynomial) crossing the threshold line compared to the reference. For the final irradiation campaign this is shown in figure 5.20. An irradiation campaign is a measurement period between two annealing processes (in case of the first campaign between the measurement start and the first annealing process). This measurement here therefore consists of six irradiation campaigns. The first campaign is shown in figure 5.23. Campaigns two to five are shown in figures 7.6 to 7.9 in the Additional measurement data section.

Frequency [GHz]	Accumulated TID above 3 dB attenuation [MGy]	Accumulated TID above 6 dB attenuation [MGy]
0.1	11.2	12.5
1	9.4	10.4
5	8.2	9.1
10	7.1	8.5
15	6.6	8.1
20	6.0	7.2
30	6.0	8.0
40	6.1	8.2

Table 5.1: Total ionizing dose durations of each frequency's electrical-optical-electrical response above the threshold values.

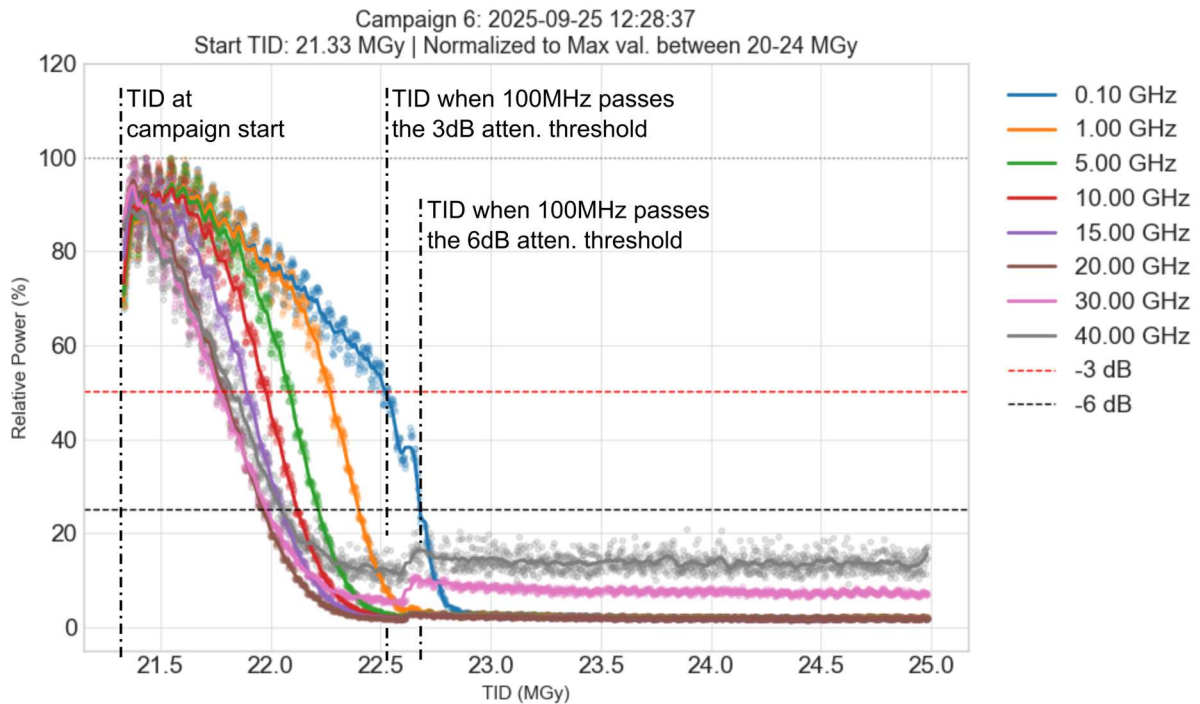


Figure 5.20: The final (sixth) irradiation campaign of the ring modulator sample. The start of the campaign and places where the 100 MHz curve passes the 3 dB and 6 dB attenuation thresholds are indicated with vertical lines. The threshold depended lifetime within a given campaign is the TID distance between the start line and the threshold lines.

A clear low-pass behavior for both frequency and modulation lifetime for progressing campaigns versus received TID becomes apparent. The reduction in lifetime of the individual campaign versus TID is

shown in figure 5.21. The curves exhibit asymptotic behavior for all frequencies with regard to the required dose until modulation collapses. The Bode diagrams (figure 5.22) of the individual campaigns also visualize this behavior. This is in line with the review [50], stating that while oxide trapped charges may anneal over time, interface states tend to be stable and do not significantly recover at room temperature. The observed low-pass behavior results from the radiation-induced increase in series resistance [51], which, combined with the constant parasitic capacitance, increases the RC time constant and thus limits the bandwidth. In other words, it introduces a frequency dependent impedance mismatch

$$Z = R_S + \frac{1}{i 2\pi f C_p}, \quad (5.1)$$

where R_S is the variable series resistance and C_p the constant parasitic capacitance. R_S cannot be extracted from this dataset, since it must either be measured in a FWB or via the S-parameter.

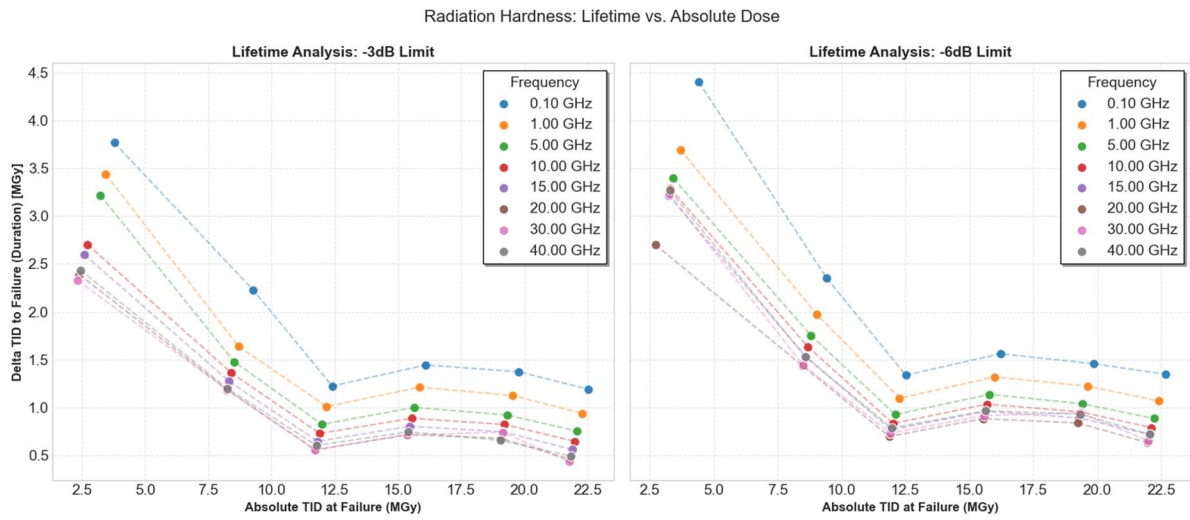


Figure 5.21: Modulation lifetime-changes for individual modulation frequencies at progressing accumulated total ionizing doses.

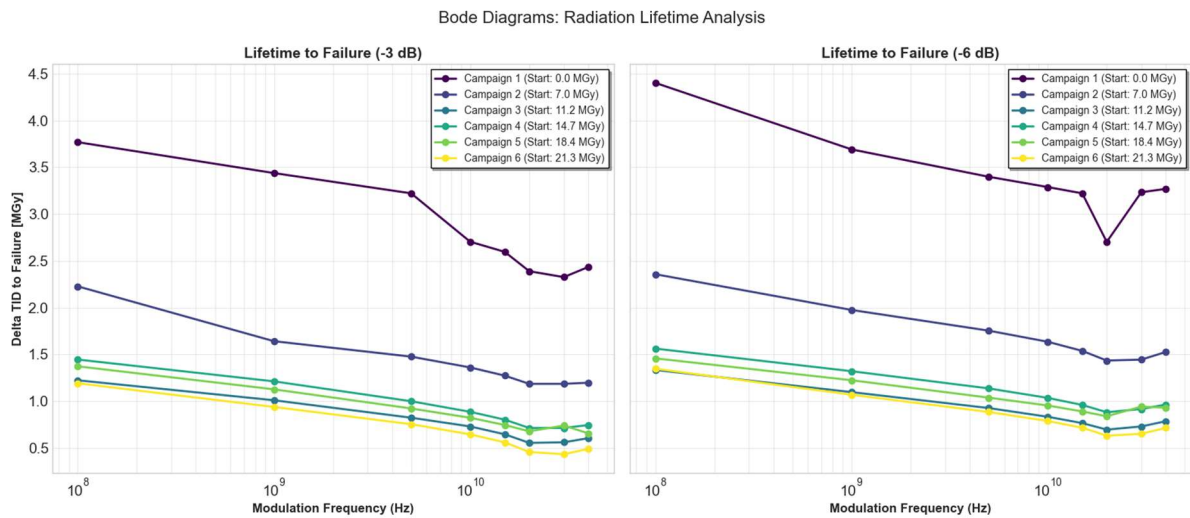


Figure 5.22: Bode diagrams of the modulation lifetime for all irradiation campaigns.

The data at 30 GHz and 40 GHz in table 5.1 suggests that these frequencies perform better judged by their lifetime compared to the next lower frequencies. This can also be seen in figures 5.21 and 5.22. However, this performance “increase” is most likely an artifact of the measurement and can be explained by the SA’s behavior for high modulation frequencies at low EOE response levels. There the SA’s inherent noise floor adds to the diminishing signal power, artificially inflating the measured EOE response. This can be seen by the offset at 5 MGy of both frequencies compared to the rest. Consequently, the simple criterion of checking when the filtered line passes the threshold values is compromised to a certain degree. Both frequencies perform a bit better. This is compounded by the large swings in the optical coupling efficiency. This and becomes apparent in the plot in figure 5.23 at approximately 3.2 MGy. Due to the lifting of the 30 GHz and 40 GHz curves the -6 dB threshold is passed again compared to the previous passing around 2.8 MGy.

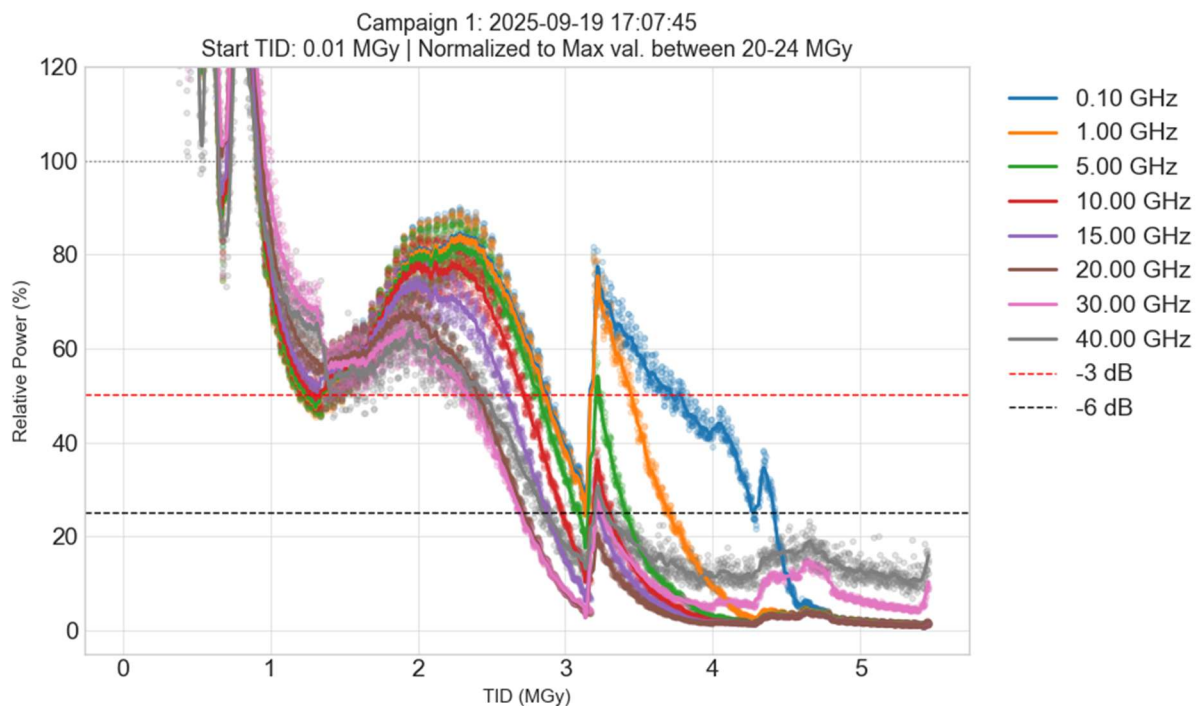


Figure 5.23: The first irradiation campaign of the ring modulator sample. The large swings in the optical coupling efficiencies influence where especially the higher modulation frequency curves pass the 3 dB and 6 dB attenuation thresholds. This influences the lifetime estimates of the ring modulator for the affected frequencies.

The analysis up to this point demonstrates the viability of RMs for in-detector use for low modulation frequencies with respect to X-ray irradiation at quasi in-situ conditions. For 100 MHz both attenuation thresholds show a lifetime over 11 MGy, which is above the requirements of the HL-LHC. Here the accumulated TIDs, however, are even higher. The asymptotic nature of the modulator lifetime across all frequencies suggests that at closer annealing intervals, also these frequencies would surpass the requirements. Additionally, in a detector the irradiation would be timed with the lifetime of the modulators used. Therefore, here a worst-case scenario is shown with regard to the radiation effects. This asymptotic behavior shows that a time-dependent annealing protocol could be formulated, ensuring continuous data transmission.

Mathias Ludwig's measurements [41] also showed that the working point shifts to higher wavelengths for higher doses. To visualize this, figure 5.24 shows the working point (laser wavelength) and the EOE response with a shared TID-axis. A possible trend towards higher wavelengths can be suggested, which is inversely correlated with the EOE response, however it would be very weak. Furthermore, the influence of the changing coupling geometry (i.e., the cause of the fluctuations) on the working point is evident, also sometimes leading the automated peak search algorithm to wrong results (changing coupling efficiency, while it is running). Therefore, ideally the notch positions should be analyzed. The working point should move exactly like the notches do, assuming there is no change in the notch shape.

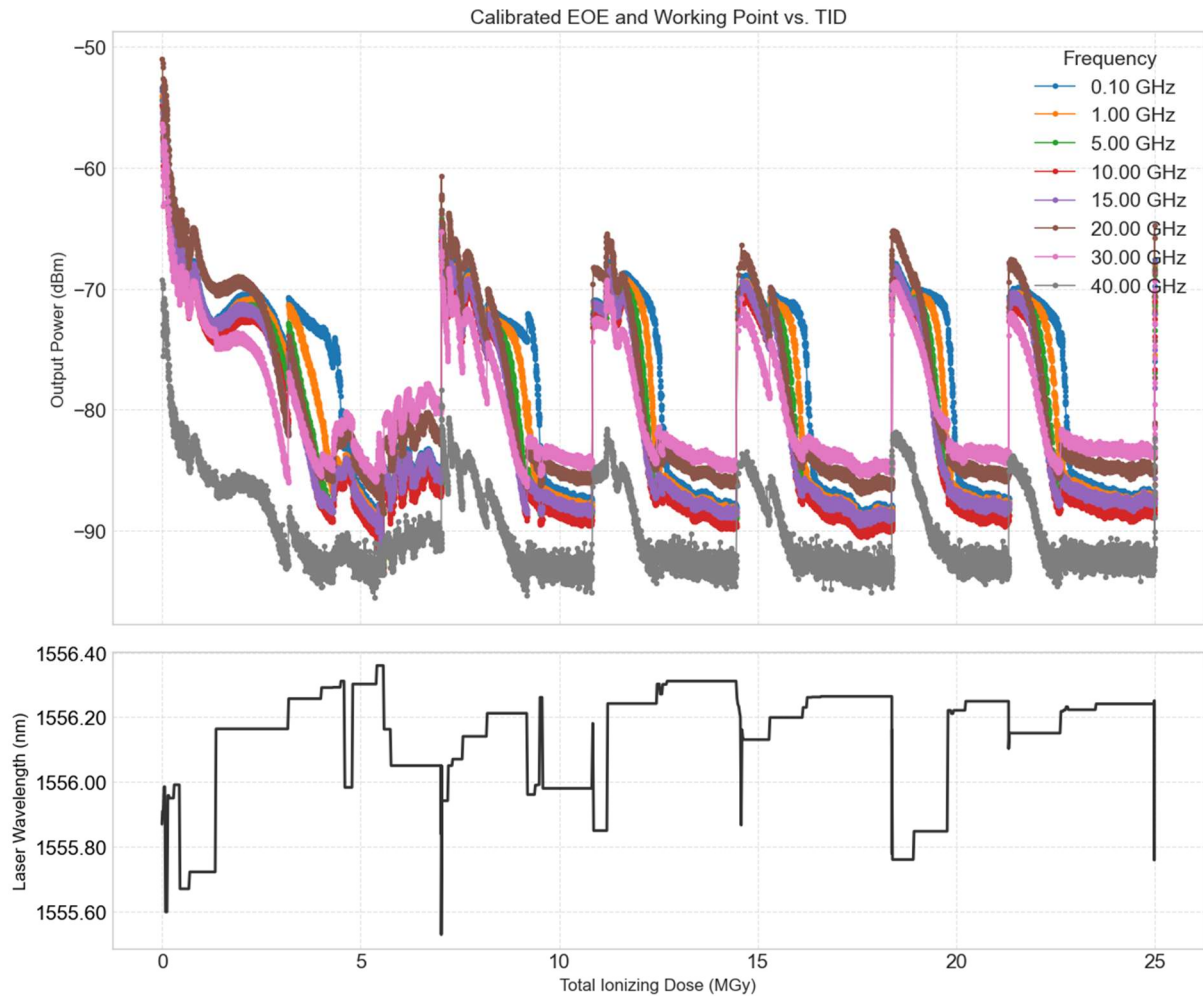


Figure 5.24: The working point (laser wavelength) of the ring modulator sample with a shared x-axis to the electrical-optical-electrical response.

Figure 5.25 shows the notch position of all three notches over the TID. The positions were derived similarly to the methodology used for the tunability investigations in section 4.3. The curves are filtered with a very wide window and then the filtered line is subtracted from the data. Then the data is flipped and fitted with a Gaussian distribution. The center of the distribution yields the peak position. However, for notches on the falling, rising or both flanks the OPM introduced plateaus due to the auto-range function. These plateaus occur relatively often and therefore have been removed. However, this decreases the available data points surrounding the notches drastically (typically from five down to four or three). This leads to higher inaccuracies in the position tracking, since the data has a resolution of 0.1 nm and just like in [41] the changes are in the sub 1 nm regime. The change in the notch position

due to TID in the plots is apparent. Especially the campaigns at high TIDs, where the swings are less pronounced, confirm this behavior. The annealing times are also recognizable. At higher TIDs the range of the spectral shift decreases to roughly 0.25 nm.



Figure 5.25: Resulting plots of the notch tracking of the DC spectra of the ring modulator sample.

To analyze this behavior further, the annealing process is closer investigated. Figure 5.26 shows the first annealing process and figure 5.27 the final (sixth) one. The remaining annealing sequences are shown in the Additional measurements data section in figures 7.10 to 7.13. The latter also include annealing processes under active irradiation. However, no difference to annealing without irradiation is discernible. The x-axis shows the total annealing time. The smallest time where FWB annealing was applied was 5 s. The left y-axis shows the absolute value of the EOE response recorded after the annealing was applied. The right y-axis shows the working point chosen by the measurement algorithm.

Above those data points the current with which was annealed is written. Only two currents were chosen: 50 μA and 200 μA . The reason to plot the working point found by the algorithm and not the notch position(s) is the low accuracy of the notch tracking. For the first annealing process, no uniform trend in the working point can be observed, as it fluctuates significantly. However, looking at all provided curves, possibly a trend towards lower wavelengths after the annealing process can be suggested. Correlating this with the EOE response at the longest annealed data points possible “over-annealing” starts, where the working point shifts notably while the EOE response degrades. Due to the limited and somewhat noisy data, this is more of a phenomenon to be investigated in future measurements. It would align with the rebound effect described in [48]. During annealing, oxide charges and interface states compete. While the application of FWB effectively heals the oxide charges, it can activate or unmask the more persistent interface states. Once these interface states dominate the device characteristics (“over-annealing”), they alter the local charge carrier concentration. Via the plasma dispersion effect (PDE), this modifies the effective refractive index, leading to the observed shift in the working point and a renewed weakening of the EOE response.

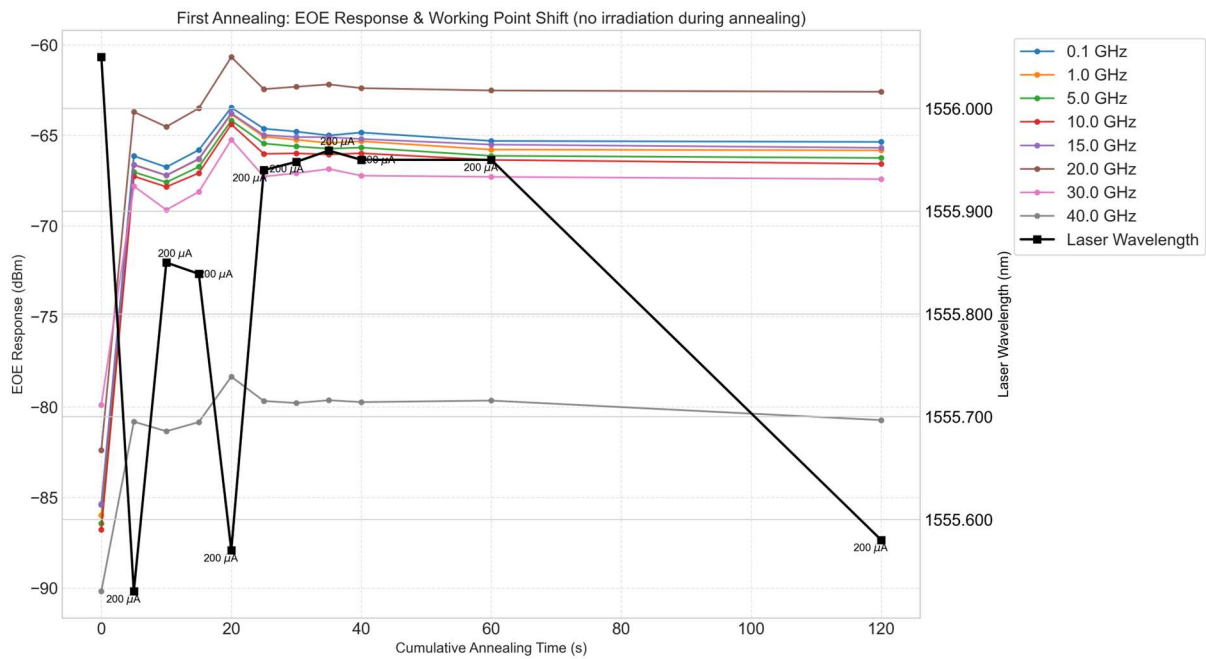


Figure 5.26: First annealing process of the ring modulator sample. All annealing steps were done with 200 μA .

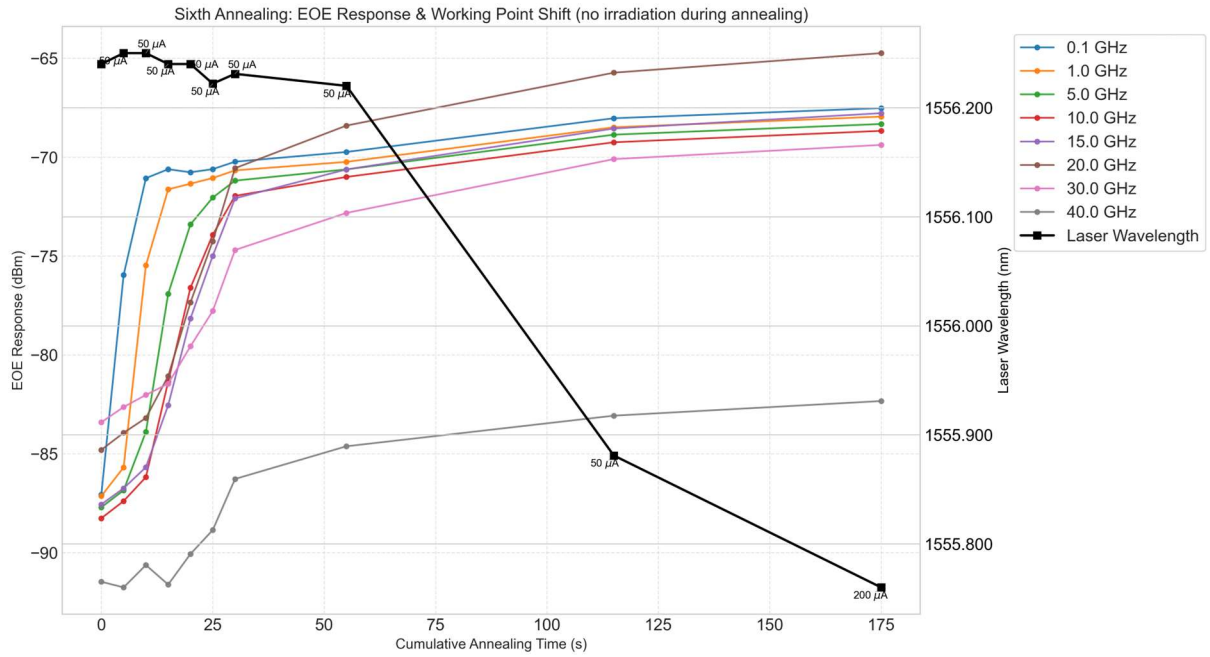


Figure 5.27: Final (sixth) annealing process of the ring modulator sample. All annealing steps were done with 50 μA except the final one which was done with 200 μA .

Although the EOE response suggests a rapid recovery in the first annealing process reaching a plateau after 25 s, the working point reveals significant volatility up until this point. Assuming this is not an artifact of the measurement algorithm it implies that the device has not reached a stable state immediately. This may hint at a transient shift of the operating point into a more efficient region, rather than solely by defect annealing.

Comparing the first and last annealing process, this transient behavior vanishes. It must be noted that in the beginning it was annealed with only 50 μA . When all annealing processes are considered, the transient behavior also seems to decrease from one annealing process to the next. The increased interface damage however shifts the point of the fully annealed state to the end or out of the recorded time frame. It becomes apparent that higher TIDs should result in higher annealing times. As mentioned previously, the mechanical instabilities of the optical coupling prohibit a direct comparison of the EOE response over the TID between different stages of the measurement.

Complementing the frequency analysis, the investigation of the working point (and notch) shift demonstrates that active working point control would be confined to a small wavelength range even at high TIDs. The very high dose rate used here is close to a worst-case prognosis in that regard. Provided that mechanical optical coupling is maintained, the required spectral tuning range is shown to be minimal (in the sub-nanometer regime). Furthermore, the annealing experiments reveal that device recovery extends beyond the mere restoration of the EOE response magnitude; it is intrinsically linked to the stability of the working point. The observed transient volatility during the annealing phases highlights that future operation protocols must account for these relaxation times to ensure not only sufficient modulation power but also a stable, drift-free locking of the operating wavelength.

Future characterizations should include two important modifications. First, after each annealing step, an ultra-short FWB IV-curve should be recorded to track the series resistance R_S of the pn-junction. This

would allow for further RF characterization regarding the dependency of frequency performance versus TID. The IV-curve parameters must be chosen carefully to not noticeably anneal the radiation damage and consequently alter other characteristics. The second change should aim to increase the accuracy of the working point or notch position determination, especially during annealing. This could be done by disabling the auto-range feature of the OPM and increasing the measurement resolution around the notches. Another approach would be to increase the data collection times of the SA to improve the algorithm's stability. Even manual optimization could yield an improvement. However, all these suggestions come with a time penalty. In light of the following MZM characterization the flat fiber-to-chip coupling without glue fixating the facet above the GC should be questioned.

5.4 Mach-Zehnder modulators

5.4.1 Deep etched Mach-Zehnder modulators

Here the irradiation experiment results of a deep (standard) etched MZM with GCPW fan-out boards (AN-2061 style bias tee and termination fan-out boards) are shown. The exact structure used is the DO5#8#L1000A50 (Wafer#Chip#Structure). The measurement lasted approximately three days, 22 hours, and 34 minutes in total, reaching an accumulated TID of 16.5 MGy. However, after reaching 12 MGy the sample wasn't irradiated for approximately 14 days. The TID was achieved with a dose rate of $48.5 \text{ Gy}\cdot\text{s}^{-1}$ again at a distance of 25 mm. The laser was set to 6 dBm. The SA's bandwidth was configured to 500 Hz, the span to 5 kHz and the reference level to -30 dBm . These values were changed compared to the RM ones, because in a test measurement, the EOE response measured by the SA were out of range. The push-pull MZM was biased at 1 V in RVB in both arms, where they share a central potential, which is modulated with the RF signal. Figure 5.28 shows all the recorded frequency data and figure 5.29 shows all DC sweep data.

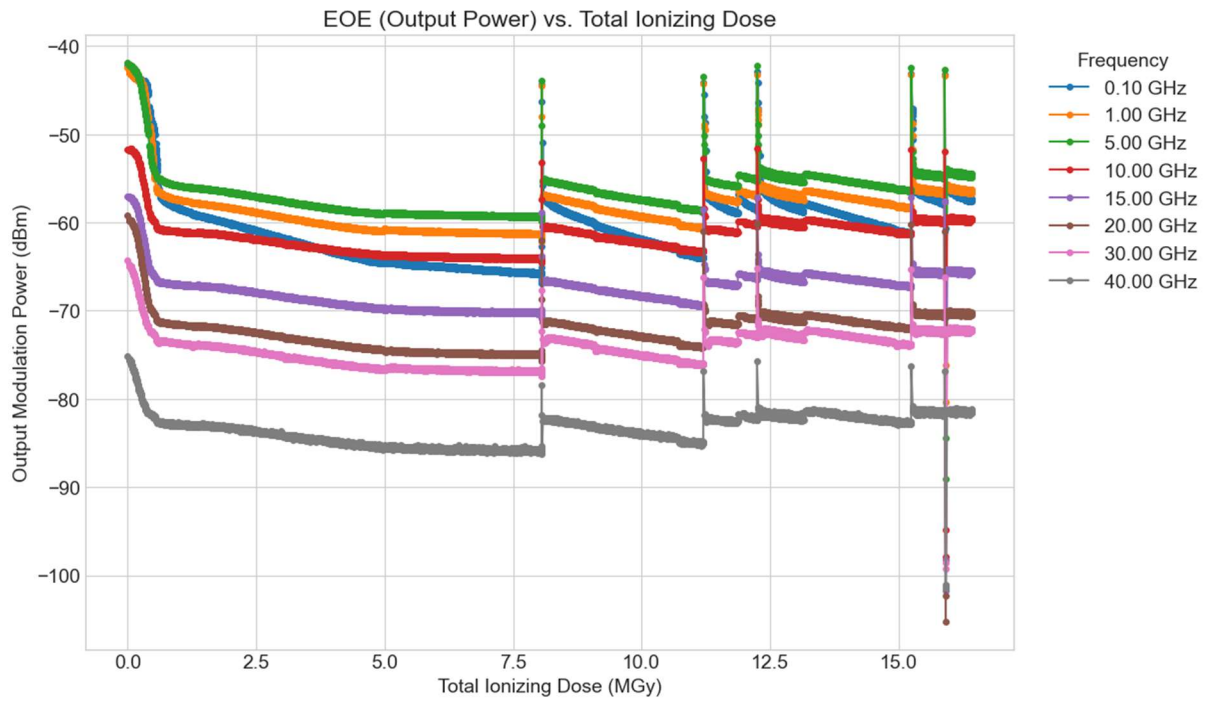


Figure 5.28: Electrical-optical-electrical response of the deep etched Mach-Zehnder modulator sample for different modulation frequencies and for an increasing total ionizing dose.

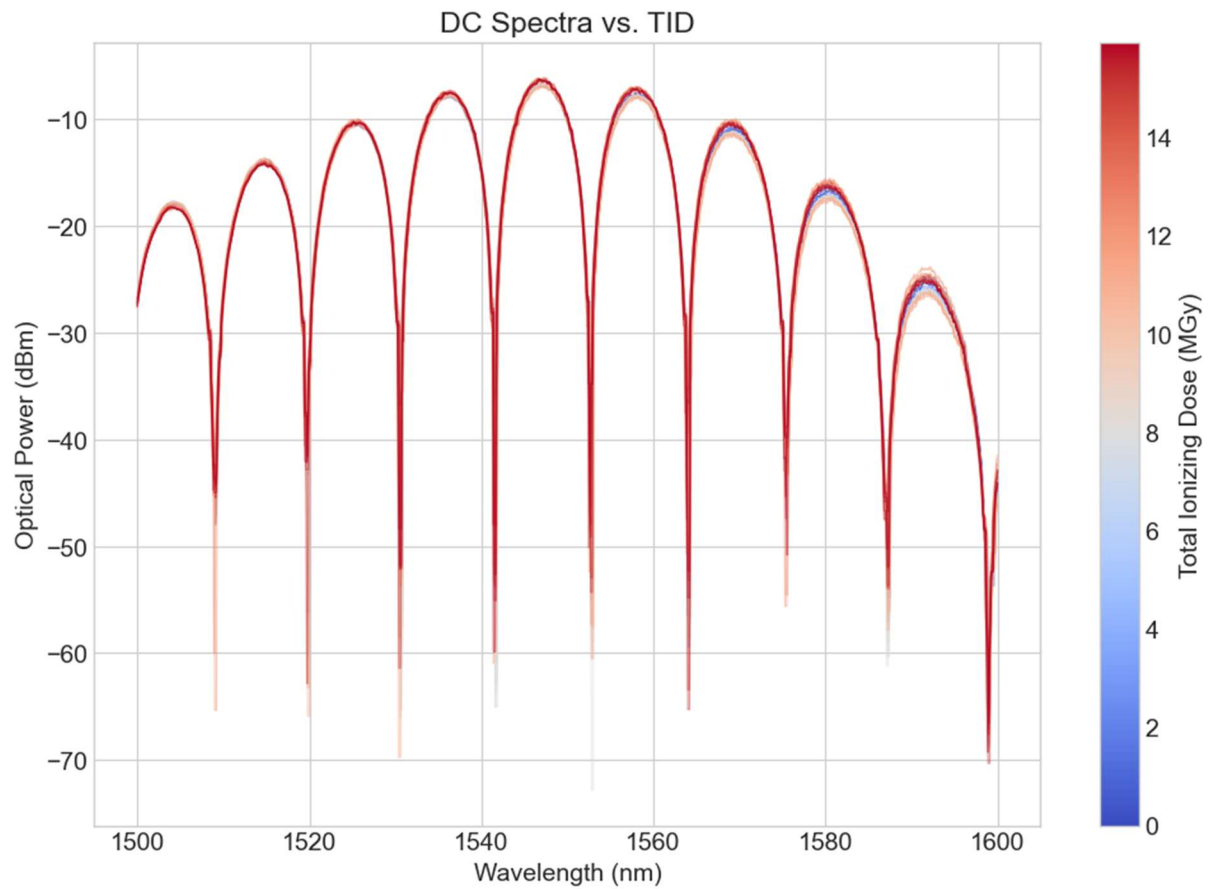


Figure 5.29: DC spectra evolution of the deep etched Mach-Zehnder modulator sample for an increasing total ionizing dose.-

The RF data initially shows a peak EOE response of roughly -42 dBm for certain frequencies. All frequency curves drop by approximately 12 dBm, where they keep deteriorating slowly. The 100 MHz curve is the exception, as it keeps deteriorating at a comparatively high rate. The 40 GHz curve here stays clear of the SA's noise floor. It was annealed five times with the same FWBs in both arms at currents ranging from 200 μ A to 50000 μ A. The timestamps (or TID levels) corresponding to the annealing processes are identifiable where the EOE response recovers from their collapsed state. No swings like in the RM measurement in section 5.3 can be observed. There the coupling efficiency was unstable especially at the beginning of the measurement leading to strong oscillations in the EOE response ("swings"). In the following the same cable compensation for the EOE response as in section 5.3 will be applied and used for all further analysis (figure 5.28 shows raw data).

The first focus here is on the DC spectra shown in figure 5.29. As already mentioned, no swings are present. Also, the coupling efficiency is much higher. This is a positive sign for the changed interconnection approach, i.e. completely fixed fibers via gluing them to the chip. Figure 5.30 shows a choice of wavelengths and their associated optical powers over the TID. The lines are remarkably straight at lower wavelengths, while at higher ones around 12 MGy jumps can be seen. This is where the measurement was paused for several days. In the meantime ambient conditions in the room have changed (temperature) and time-based annealing has occurred. Looking at the EOE response at the same TID also shows a minor jump for all frequencies, confirming the time-based annealing hypothesis. The tracking of the maximum of the filtered DC spectra curves (Savitzky-Golay with a window of 10 data points and second order polynomial) close to 1547 nm over the TID leads to the plot in figure 5.31. A relation to the working point is not as direct as with the RM (due to the higher distance between DC peak and working point for the MZM compared to notch position and working point for the RM), however, should describe trends similarly. The annealing times and the long measurement brake are directly discernable by the jumps. However, a pattern beyond this is not. Possibly a long-term shift to shorter wavelengths. The total range of the maximum position is confined to 1.1 nm.

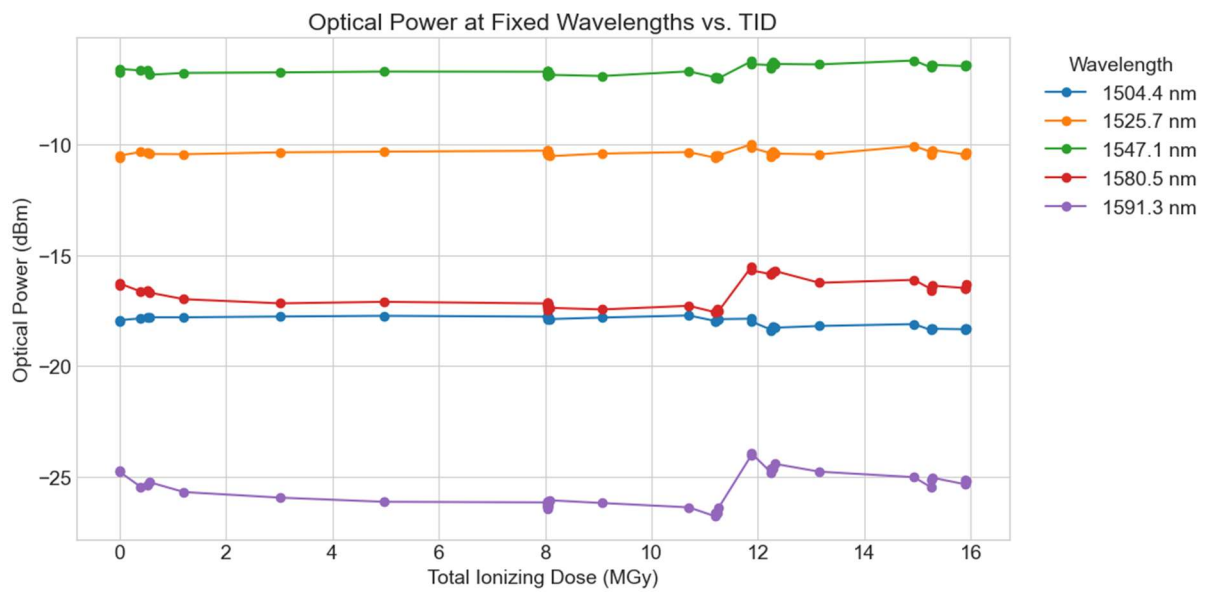


Figure 5.30: Optical power of the deep etched Mach-Zehnder modulator sample for certain wavelengths close to/on selected peaks in the DC spectrum at increasing total ionizing doses.

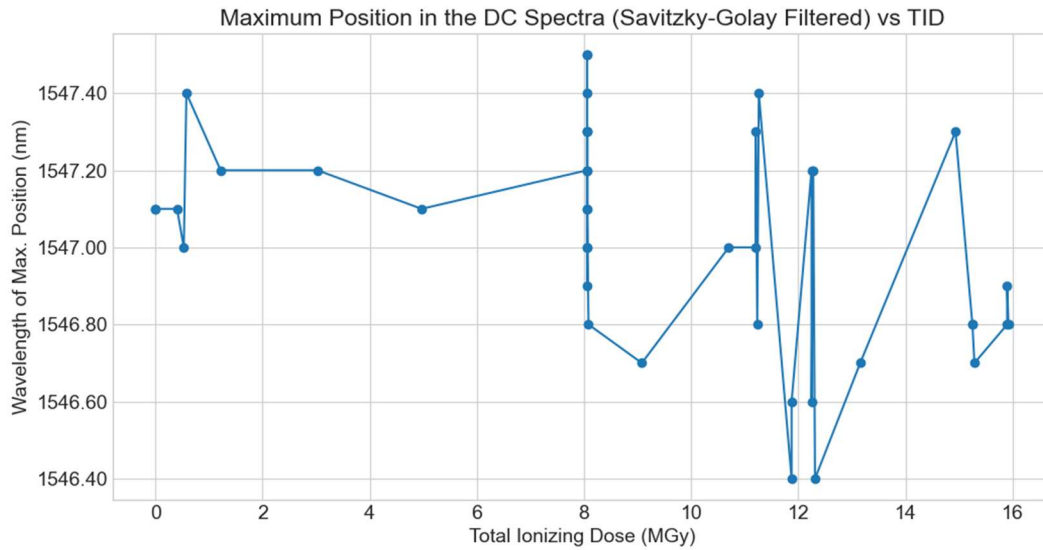


Figure 5.31: Tracking of the maximum position of the (filtered) DC spectra of the deep etched Mach-Zehnder modulator sample at increasing total ionizing doses.

Figure 5.32 shows the now corrected EOE dataset and the working point determined by the algorithm. Directly apparent is a problem of the algorithm to choose which slope of a DC spectrum to modulate on. It jumps between the rising and falling slope of the “peak” around 1547 nm. Additionally, the very short lifetime of the deep etched MZM is obvious. The annealing processes don’t revive the modulator long enough to do a representative lifetime or frequency behavior analysis. Figure 5.33 contains the first irradiation campaign and figure 5.34 the fifth (a campaign is the period in the measurement between the annealing processes). The other campaigns are shown in the Additional measurement data section in figure 7.14 to 7.17. The 100% reference point for each frequency is its first unirradiated single data point. While the first campaign is usable for further analysis, all others contain only the annealed data point in the beginning and an (almost) completely decayed one after the DC has been recorded and the working point has concluded. The further degradation of the 100 MHz even after the collapse may again be explained by the increased series resistance due to carrier mobility degradation induced by interface traps [50]. This parasitic resistance prevents sufficient charge carrier modulation for the plasma dispersion effect (PDE), as a large part of the signal is lost in the passive contact regions. Generally, the frequency response of this sample prefers 5 GHz hinting to good transmission characteristics for this frequency.

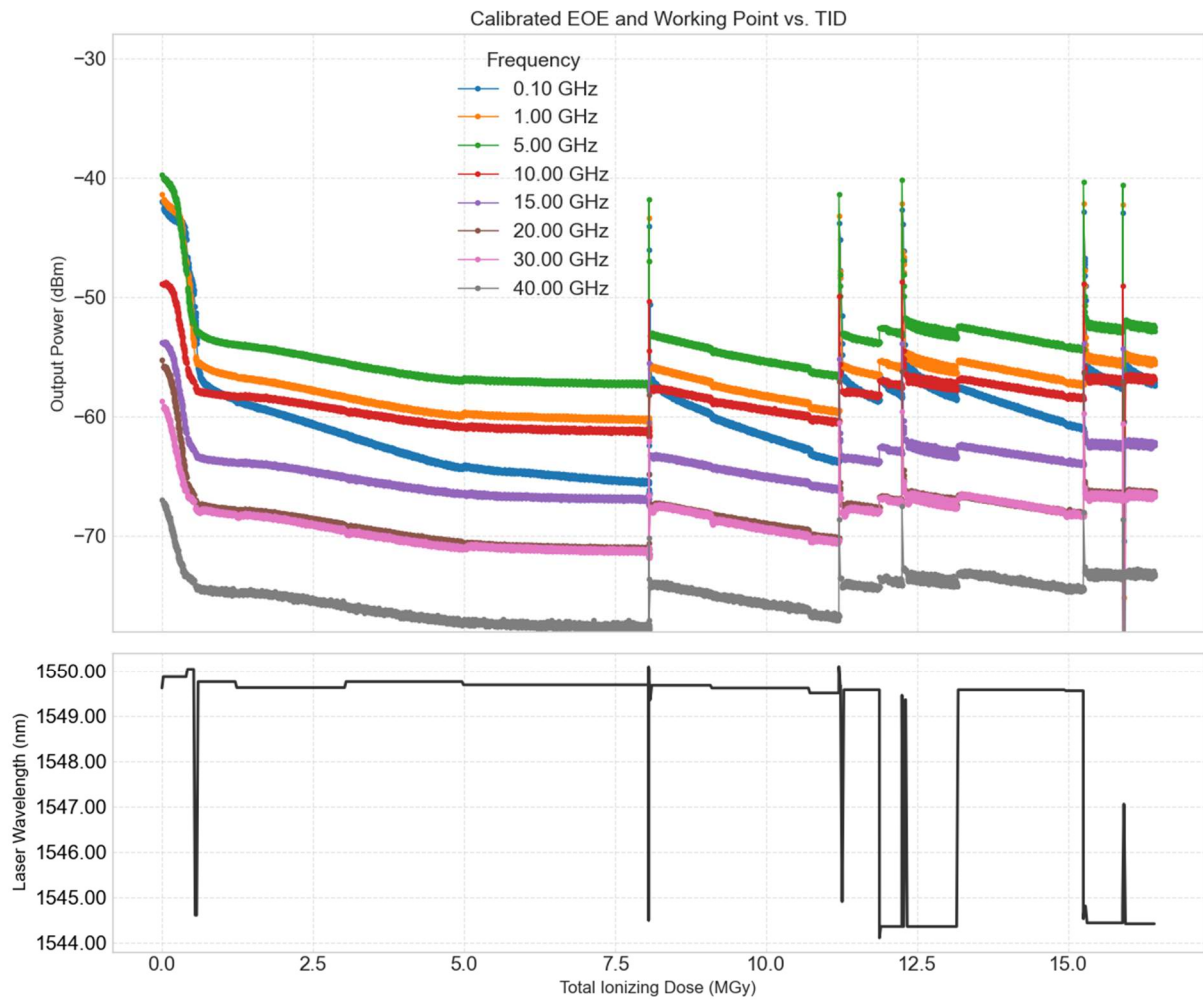


Figure 5.32: The working point (laser wavelength) of the deep etched Mach-Zehnder modulator sample with a shared x-axis to the electrical-optical-electrical response.

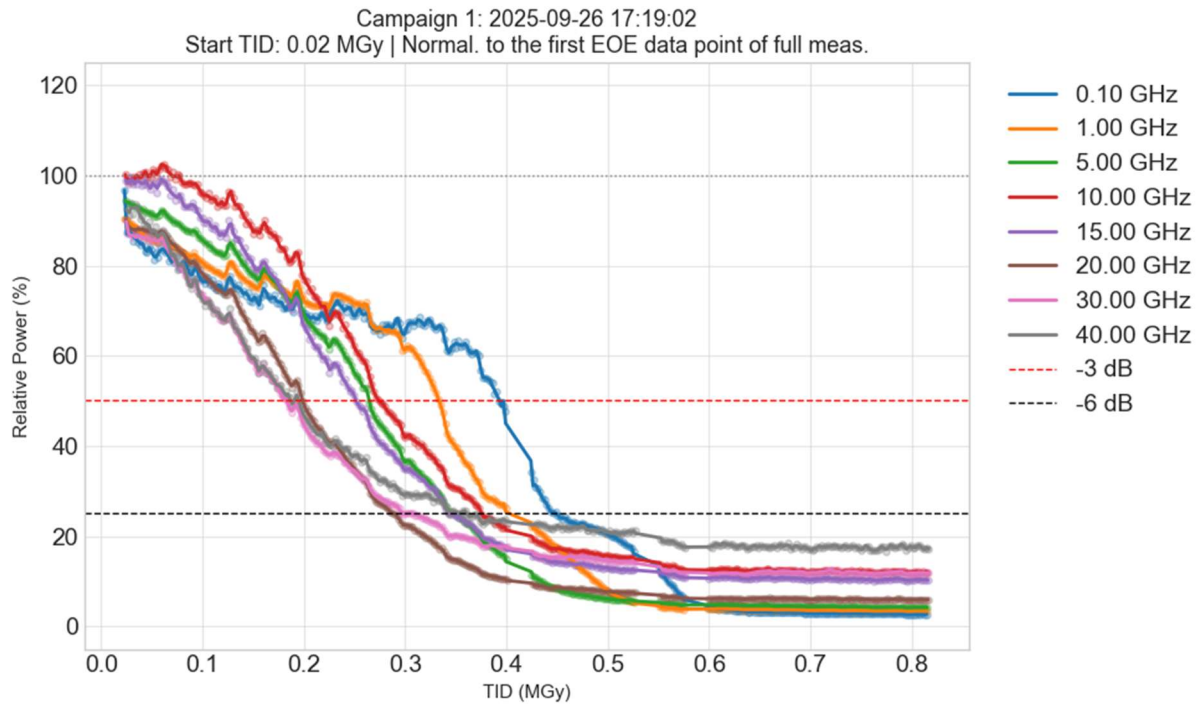


Figure 5.33: The first irradiation campaign of the deep etched Mach-Zehnder modulator sample.

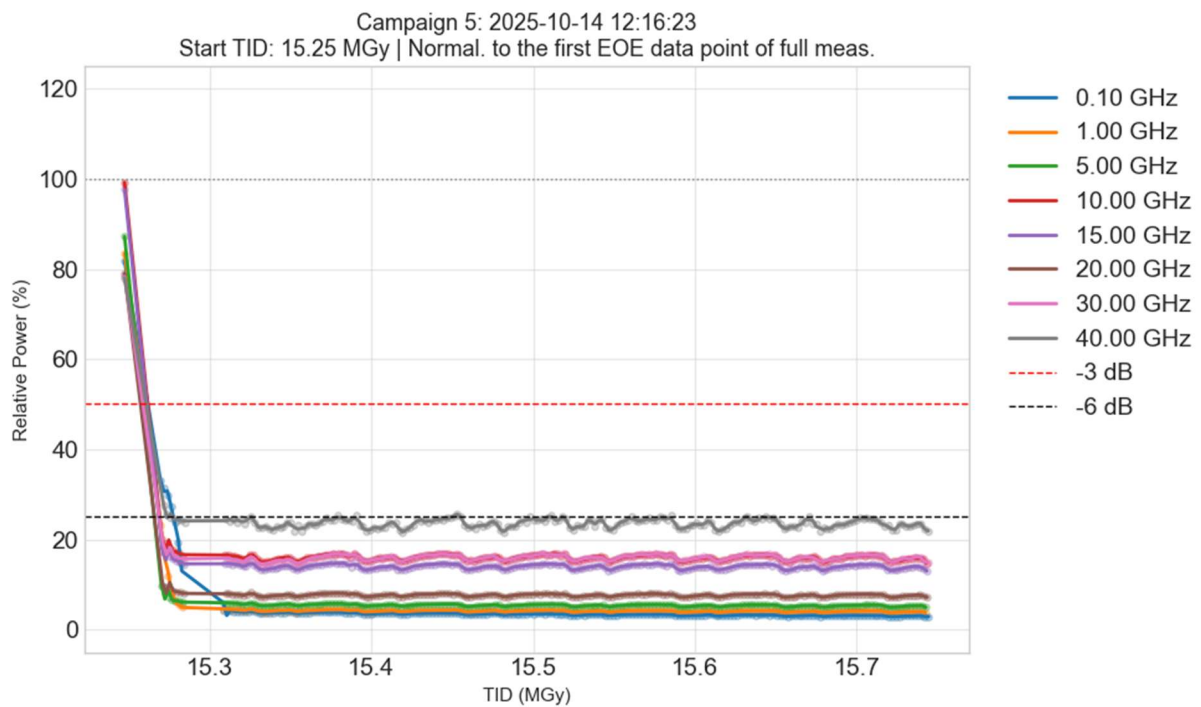


Figure 5.34: The fifth irradiation campaign of the deep etched Mach-Zehnder modulator sample.

The first annealing process of the deep etched MZM is shown in figure 5.35. Since the annealing process is strongly dependent on the electron density in the damaged region, the plot shows the higher currents needed for these large MZMs (compared to the RM size) to anneal the irradiation damage. This is so that a similar current density can be achieved. The other annealing steps are shown in the Additional measurement data section in figure 7.18 to 7.21. It was annealed with currents up to 50000 μA . Judging

only by the annealing processes alone, it appears as if with enough current density a deep etched MZM can be revived. The minimum current for reasonable annealing times (low single digit minute range) would be around 1000 μA . As apparent from the EOE response development seen above after a annealing process the modulator “dies” directly.

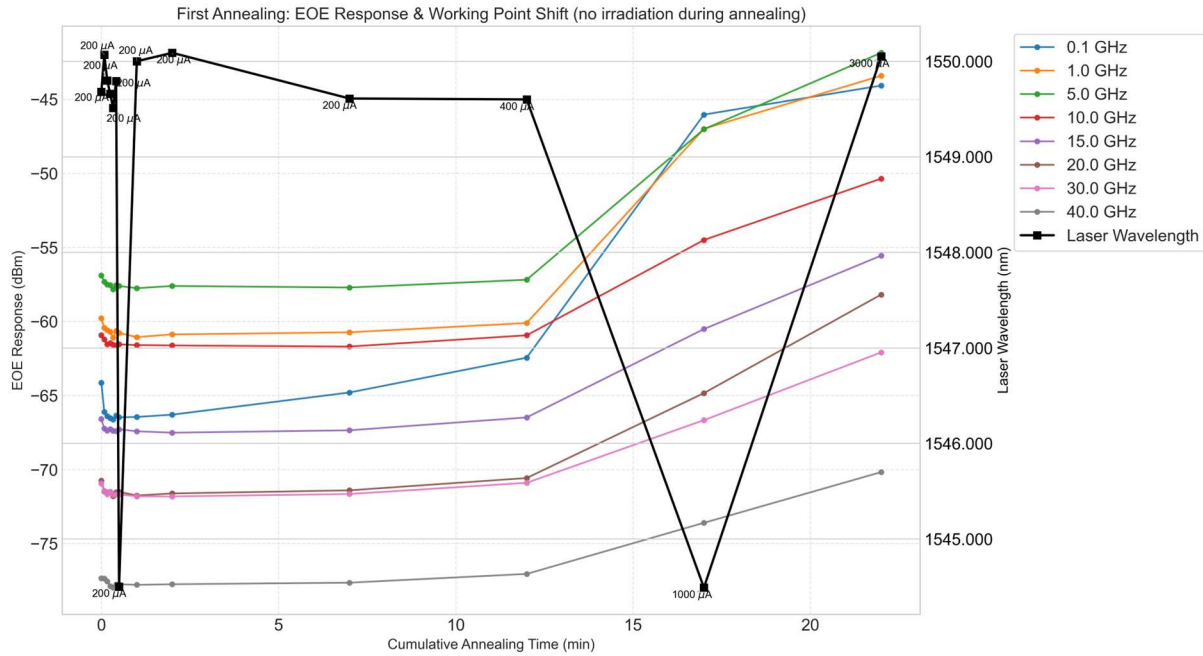


Figure 5.35: First annealing process of the deep etched Mach-Zehnder modulator sample. In the beginning all annealing steps were done with 200 μA . This was then increased step by step up to 3000 μA .

Even if the recovery is very short lived, the level of recovery with regard to the unirradiated condition is of interest to judge long-term degradation. The plot in figure 5.36 shows this, by plotting the relative EOE response of the final annealing step of each annealing process with regard to the unannealed data point (the very first in the measurement). The first and second processes had too low current densities (and possibly for too short intervals, but the current density is the dominant factor). The final two show possible over-annealing, rather than a general EOE performance loss due to irradiation. The data point above 100 % is due to noise and the fact that all points are derived from a sole data point after annealing. Therefore, for TIDs up to 16 MGy the modulator does not show a meaningful maximal possible (short-lived) EOE response.

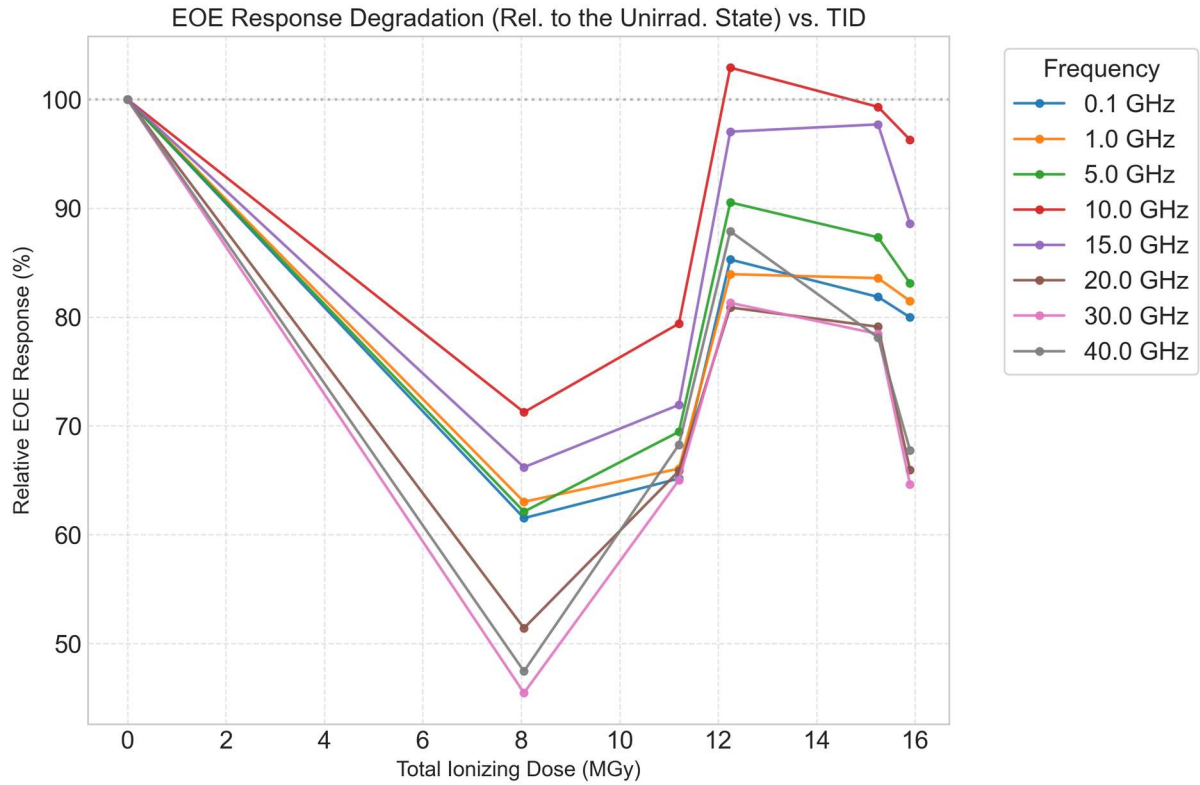


Figure 5.36: Relative electrical-optical-electrical response development after each annealing process compared to the first unannealed data points. Also the working point development is shown.

These measurements and their analysis show clearly why MZMs need to be designed with radiation hardness in mind, if their use-case requires it, even if annealing would be applied. On the other hand, it could be proven that the flat fiber-to-chip coupling works very well together with the glue fixating the angle-polished facet over the GC. If it shows the same behavior under thermal stress and neutron irradiation, the deployment of the flat fiber-to-chip coupling within a detector is a realistic scenario.

5.4.2 Shallow etched Mach-Zehnder modulators

Next the irradiation experimental results of a shallow etched, radiation hardened MZM sample are discussed. The measurement was performed on the device DO5#9#L1000A50SE (Wafer#Chip#Structure) with GCPW fan-out boards (AN-2061 style bias tee and termination fan-out boards). Since at the end of this subsection a second sample will be introduced, the former will be referred to as the first sample. The measurement lasted approximately two days, 16 hours and 57 minutes in total, reaching an accumulated TID of 10.2 MGy. The TID was achieved with a dose rate of $48.5 \text{ Gy} \cdot \text{s}^{-1} \cdot 0.9 = 43.7 \text{ Gy} \cdot \text{s}^{-1}$ at a distance of 25 mm. This is identical to the RM in section 5.3. All settings of the setup are identical to the ones of the previous subsection 5.4.1. Figure 5.37 shows all the recorded frequency data and figure 5.38 shows all DC sweep data.

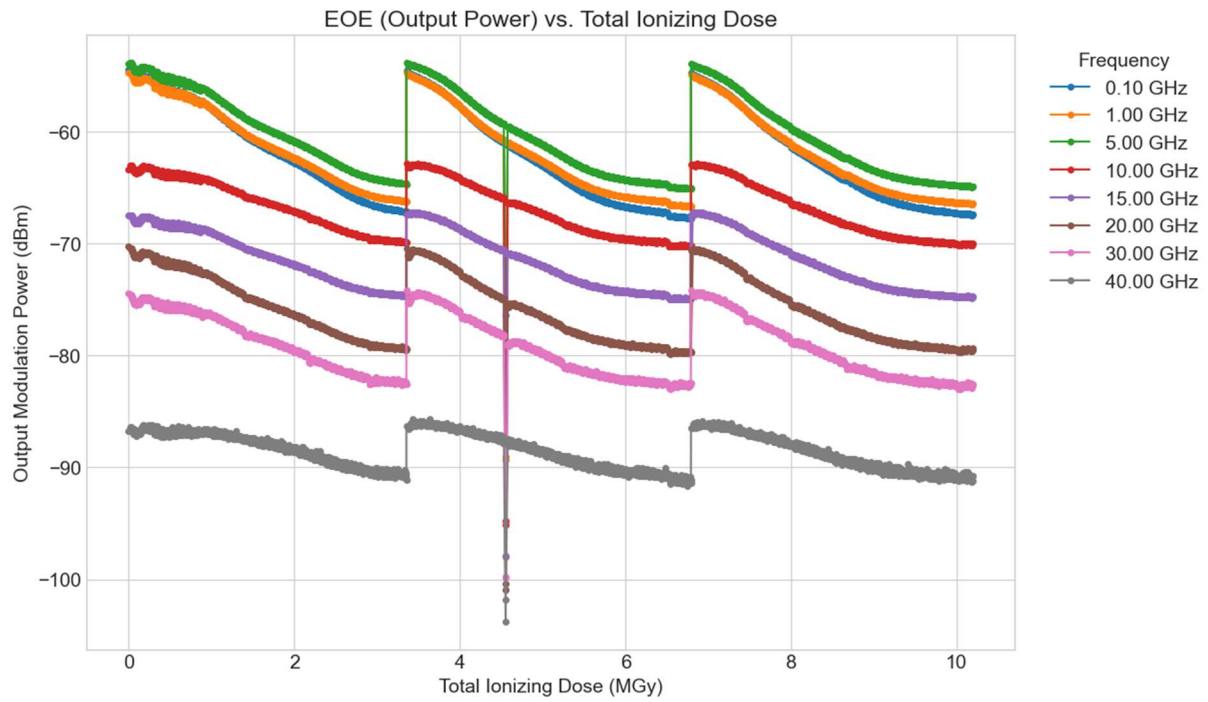


Figure 5.37: Electrical-optical-electrical response of the first shallow etched Mach-Zehnder modulator sample for different modulation frequencies and for increasing total ionizing dose.

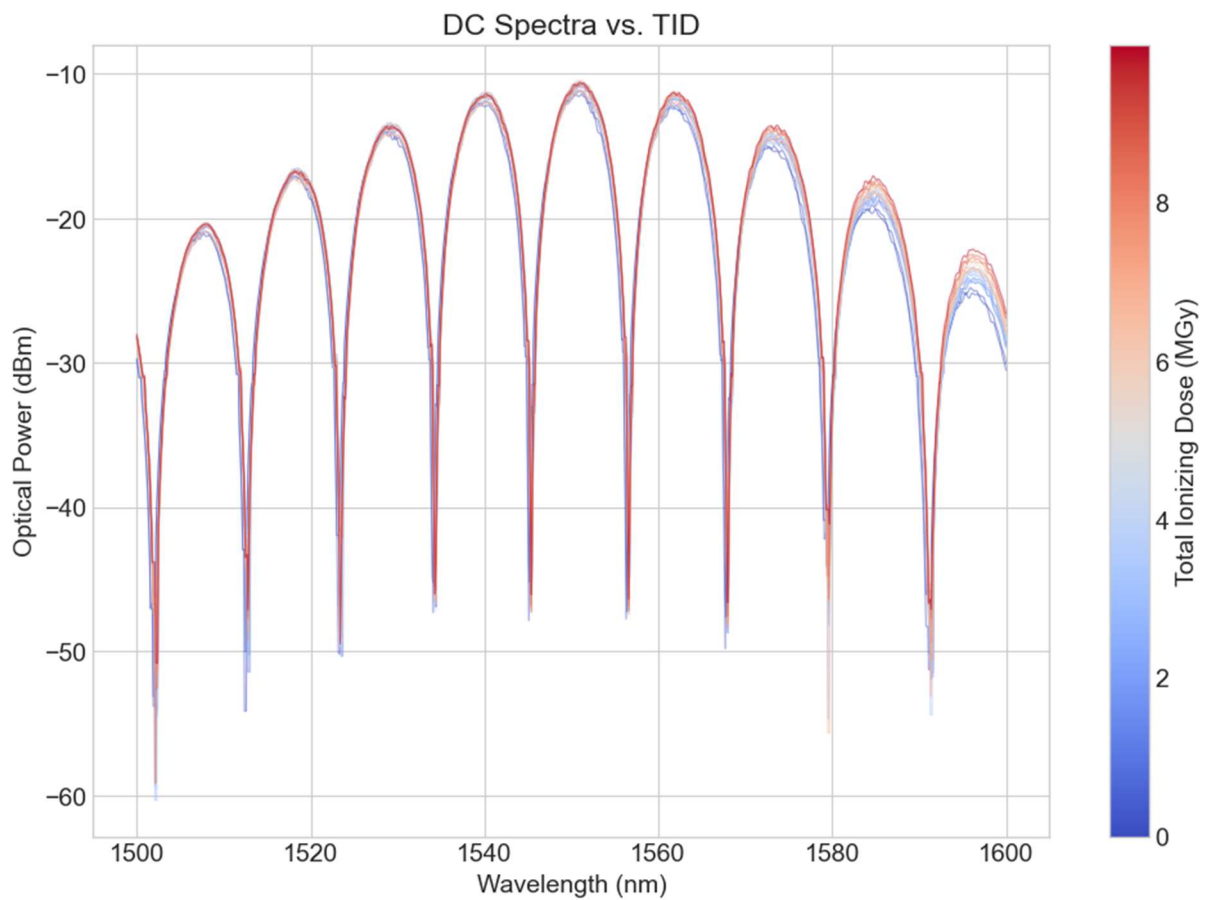


Figure 5.38: DC spectra evolution of the first shallow etched Mach-Zehnder modulator sample for an increasing total ionizing dose.

The RF data initially shows a peak EOE response of roughly -55 dBm for certain frequencies. All frequency curves drop by approximately 10 dBm through one irradiation campaign. The 100 MHz curve like for the deep etched MZM in subsection 5.4.1 keeps deteriorating after the modulation collapse at a comparatively high rate. The 40 GHz curve comes close to the SA's noise floor but never reaches it. FWB annealing was applied twice. Both times it was annealed with $35000\text{ }\mu\text{A}$ for 120 s. The annealing occurs where the EOE response recovers from their collapsed state. The cable compensation is also applied for all following EOE analyses.

First the DC measurements will be discussed. In comparison to the deep etched MZM of subsection 5.4.1, generally lower optical power was measured. This can be due to sample build variances, but can generally be expected for radiation hardened MZMs, because of their higher absorption due to doping and worse mode field confinement. Figure 5.39 shows a choice of wavelengths and their associated optical powers over the TID. The lines are very straight, which means the coupling is again very stable. Some lines show a trend to higher coupling efficiencies, especially the 1596 nm one. This can be explained by the OPM having a higher susceptibility to thermal noise at these wavelengths. During this measurement the room got warmer due to general weather trends. Also, some devices were exposed to sunlight at times. For figure 5.40 the same maximum tracking analysis from 5.4.1 was utilized. The maximum position here is close to 1550 nm, which is the desired value, and is locked to a 0.5 nm range. This means that the real best working point moves in a similarly slim range. No other trend can be derived from this curve. The fact that the maximum is at 1550 nm by chance the two angle-polished glued down fibers together produce the desired spectrum. The chuck version with the full aluminum parallelogram shown on the right in figure 4.16 was used. As stated in section 4.2 this version didn't undergo the angle optimization process. Judging by the spectra here, the optimal angle (or angle combination) was achieved by chance.

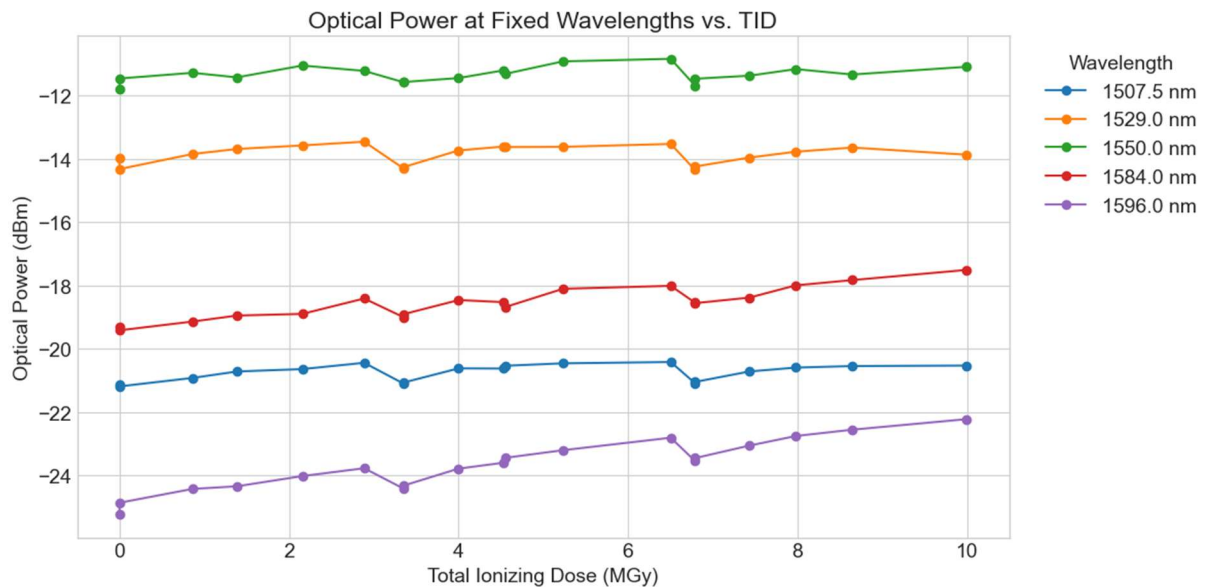


Figure 5.39: Optical power of the first shallow etched Mach-Zehnder modulator sample for certain wavelengths close to/on selected peaks in the DC spectrum at increasing total ionizing doses.

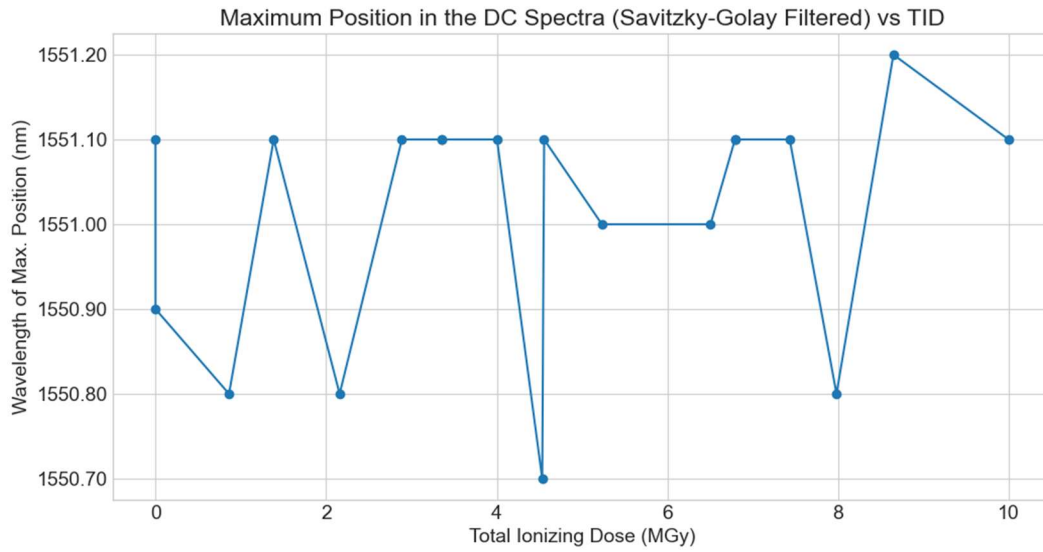


Figure 5.40: Tracking of the maximum position of the (filtered) DC spectra of the first shallow etched Mach-Zehnder modulator sample at increasing total ionizing doses.

Figures 5.41 to 5.43 show when the single filtered frequency curves pass the -3 dB and -6 dB thresholds. The analysis is similar to the one from section 5.3 for the RM. In contrast to that analysis, here the first recorded data point for any curve was taken as the 100 % reference value, because the optical coupling can be assumed to be stable and these are the only data points taken without active and previous irradiation. If the first point of the filtered line is taken as reference the threshold passes would shift, but the general trends remain identical. However, the 40 GHz line here is unlikely high. This is most likely due to three factors: The noise, generating a very low first data point, the proximity to the noise floor of the SA possibly pushing the values, due to the higher span value chosen and finally the cable correction compounding the previous points. For further analysis the results derived from the 40 GHz curve are plotted where possible, but the false shift to higher lifetimes needs to be kept in mind. The first irradiation campaign shown in figure 5.41 has some strong oscillations in the beginning, the source of this behavior is not known. Generally, all EOE response curves of all investigated modulators show some degree of these permanent oscillations. Due to the high data point density, the rapid oscillations are visually merged, causing the curves to just look broader.

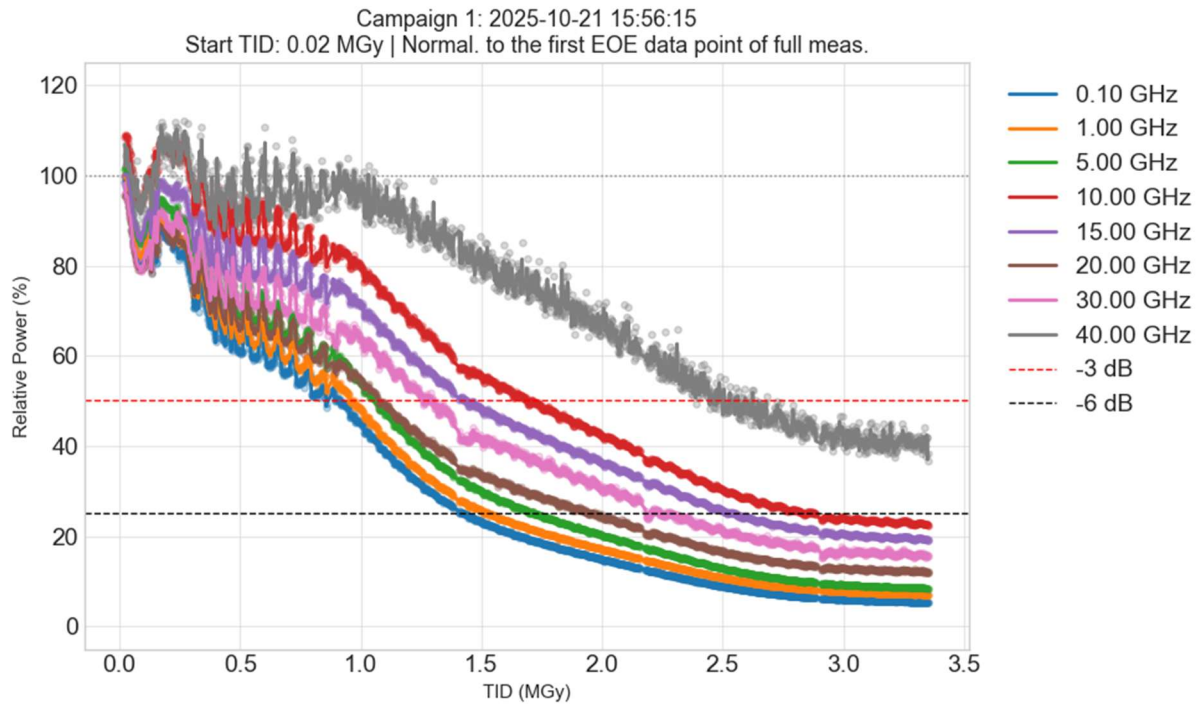


Figure 5.41: The first irradiation campaign of the first shallow etched Mach-Zehnder modulator sample.

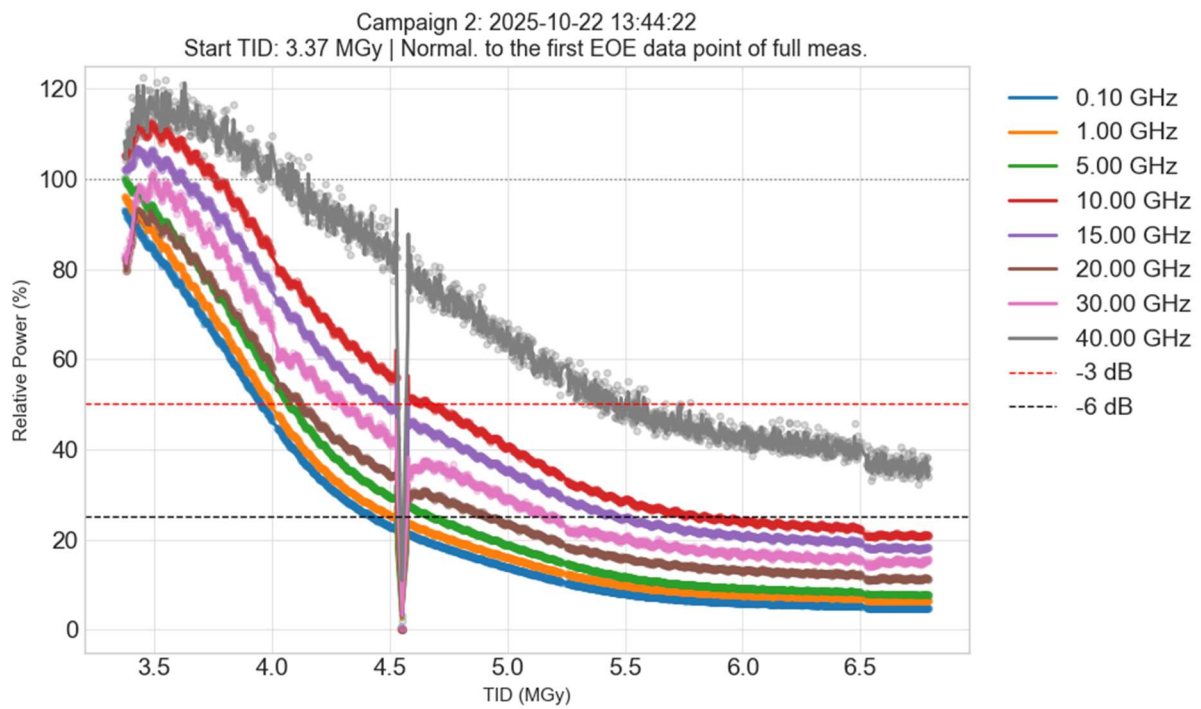


Figure 5.42: The second irradiation campaign of the first shallow etched Mach-Zehnder modulator sample.

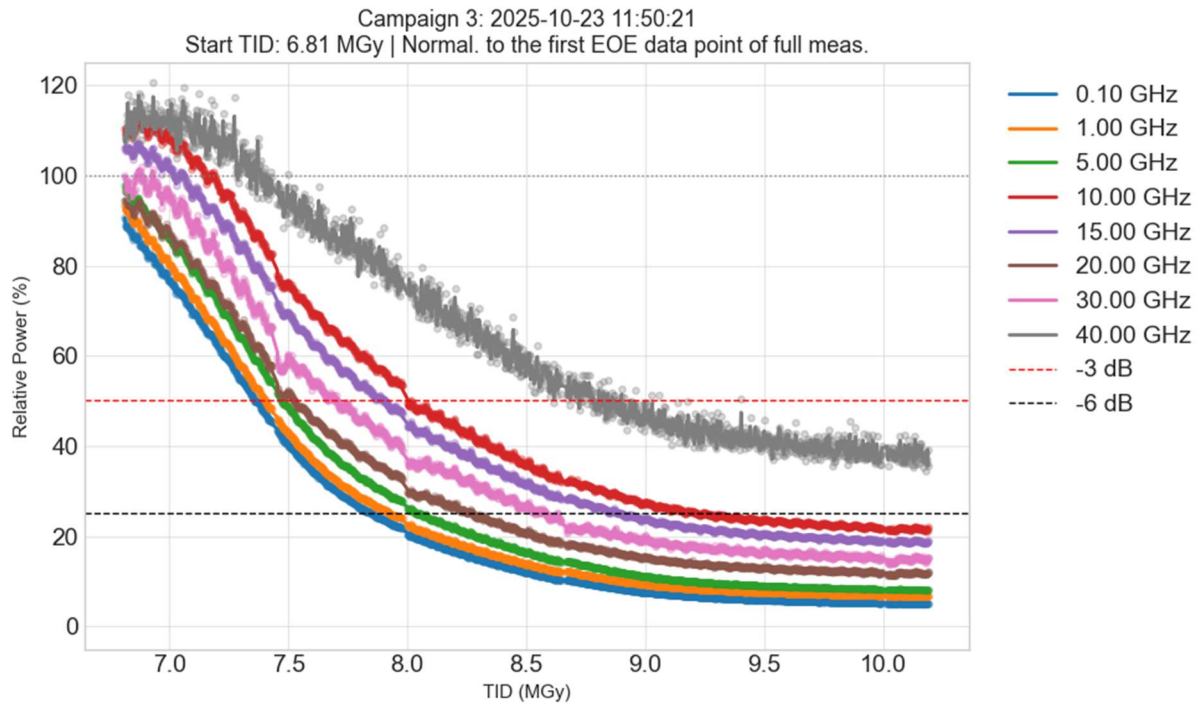


Figure 5.43: The final (third) irradiation campaign of the first shallow etched Mach-Zehnder modulator sample.

Just like the RM in section 5.3 this shallow etched MZM shows a low-pass behavior in its modulation lifetime progression versus TID received. The reduction in lifetime of the individual campaign versus TID is shown in figure 5.44. The curves exhibit a similar asymptotic behavior as for the RM. This is a strong indicator that the radiation hardened MZMs can also withstand very high TID like the ones necessary for the HL-LHC upgrade. Concerning the modulation frequency response, the Bode diagrams in figure 5.45 show the asymptotic behavior of the MZM but the frequency performance is somewhat flipped. Looking at figures 5.37 and 5.41 this result becomes apparent. Low frequencies start to decline quicker and further compared to higher ones. As mentioned above, the effect of the offset together with the proximity of especially the 40 GHz to the SA's noise floor increases this effect. However, 10 GHz and 15 GHz are the next best performing frequencies. This behavior cannot be explained by influence of the cable or the fan-out PCB since here the relative EOE response is shown. Looking at figure 5.33 showing the first irradiation campaign of the deep etched MZM sample, it shows the same frequency response with regards to lifetime as the RM. This means, lower frequencies go hand in hand with higher lifetime. So the behavior of the shallow etched MZM might be caused by the lower mode confinement and different electrical field/charge distribution together with longer current injection at low frequencies (the potential of the RF signal stays low or high longer compared low frequencies).

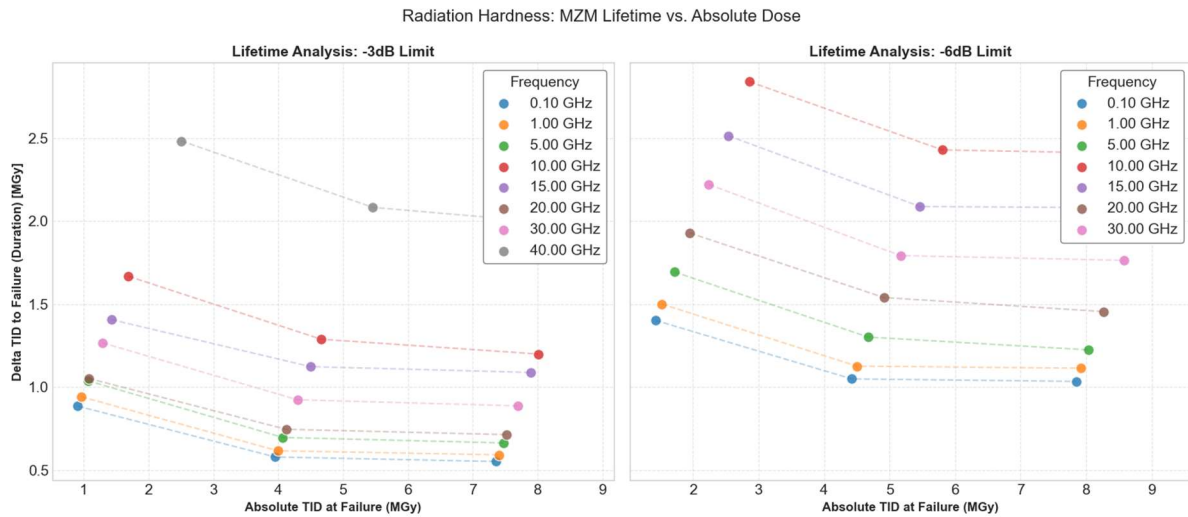


Figure 5.44: Modulation lifetime-changes for individual modulation frequencies at progressing accumulated total ionizing doses of the first shallow etched Mach-Zehnder modulator sample.

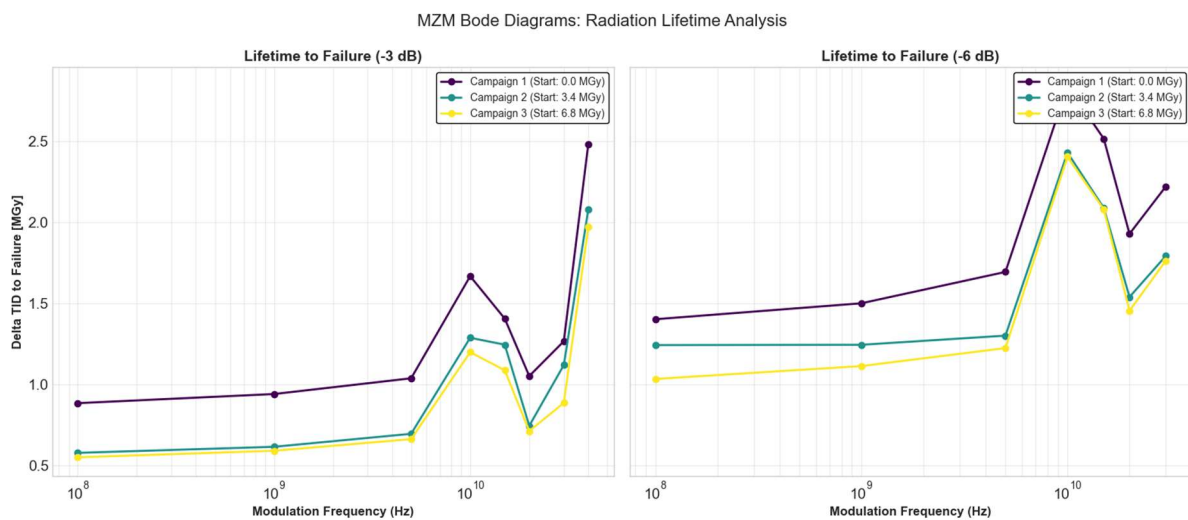


Figure 5.45: Bode diagrams of the modulation lifetime for all irradiation campaigns of the first shallow etched Mach-Zehnder modulator sample.

Continuing with the annealing and further working point analysis figure 5.46 shows the working point and the EOE response with a shared x-axis. Perhaps within a campaign a slight increase in the working point can be seen. The same instability as from the deep etched MZM measurement occurs again, where the algorithm choses a working point on the opposing flank of the 1550 nm peak in the DC spectrum. Generally, the data is just not well resolved enough and suffers from too much noise to argue for or against a relation between TID and working point. As previously discussed, the peak shifts within a 0.5 nm window. To resolve a possible link, a higher wavelength resolution than 0.1 nm is necessary.

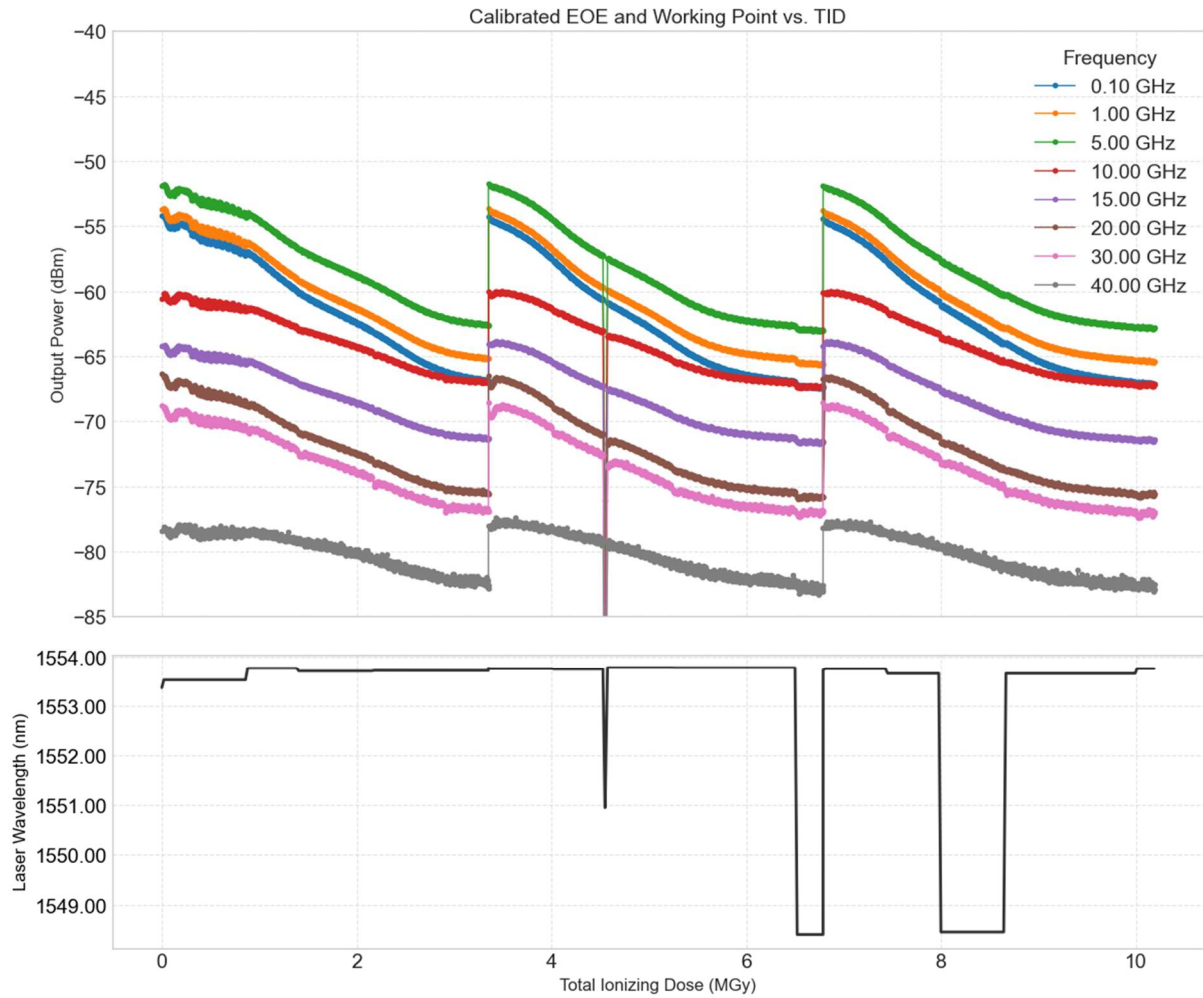


Figure 5.46: The working point (laser wavelength) of the first shallow etched Mach-Zehnder modulator sample with a shared x-axis to the electrical-optical-electrical response.

Figures 7.22 and 7.23 in the attachment show the two annealing steps undertaken for this measurement. Beyond the recovery, their meaning is limited. Figure 5.47 shows the relative change of each frequency directly after a given annealing processes has concluded. The EOE response here is in relation to the first unirradiated data point (first data point of the measurement) for each frequency. The fan-out behavior shows a similar frequency dependence as the lifetime does. For the second process close to 7 MGy all but one curve drop, showing a potential long-term degradation. This could be due to the effect observed on the RM in section 5.3. There at higher TIDs more FWB current (density) and annealing time were needed to achieve full recovery.

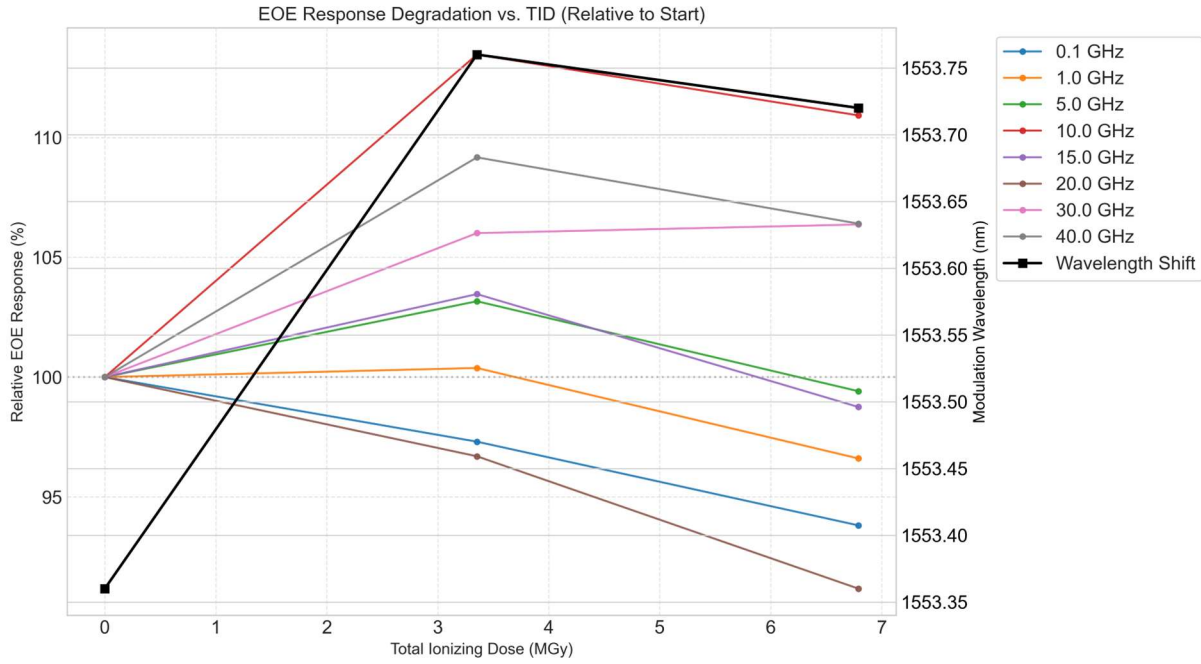


Figure 5.47: Relative electrical-optical-electrical response development after each annealing process compared to the first unannealed data points. Also the working point development is shown.

In conclusion it could be again shown that the flat fiber-to-chip coupling works very well with the glue fixated fibers. Also, radiation hardened MZMs are most likely viable for very high TID environments, but at the cost of driver electronics, which can withstand the higher annealing current requirements. The impression that a modulator that has already experienced high accumulated TIDs needs stronger or longer annealing to completely recover its modulation functionality was fortified.

It could be shown that radiation hardened MZMs are viable for the use in quasi-detector environments even for higher frequencies. The findings show that methods to closer investigate frequency behavior with regard to the TID should be conceived. For this especially the series resistances of the devices should be of interest, since they dominate the behavior at lower frequencies, as discussed in the RM section 5.3. A working point shift with changing TIDs could not be confirmed. In a detector, the challenge for active working point control will most likely stem from changing temperatures induced by X-ray or particle irradiation, rather than its primary effects on the modulator. However, for room temperature measurements, this is not an issue. The frequency response in terms of lifetime of shallow etched MZMs should also be investigated further.

A second shallow etched MZM sample measurement was performed. The device DO5#6#L1000A50SE (Wafer#Chip#Structure) was used and also connected via an AN-2061 inspired fan-out GCPW PCB (the termination fan-out PCB was also a GCPW design). The measurement's EOE response is presented in figure 5.48. The measurement lasted approximately three days, 23 hours, and 45 minutes in total, reaching an accumulated TID of 14.2 MGy. The first part until approximately 3 MGy was measured with a forward bias (FWB) instead of the standard RVB. The TID was achieved with a dose rate of $48.5 \text{ Gy} \cdot \text{s}^{-1} \cdot 0.85 = 41.2 \text{ Gy} \cdot \text{s}^{-1}$ at a distance of 25 mm. All instruments are configured like in the RM measurement in section 5.3.

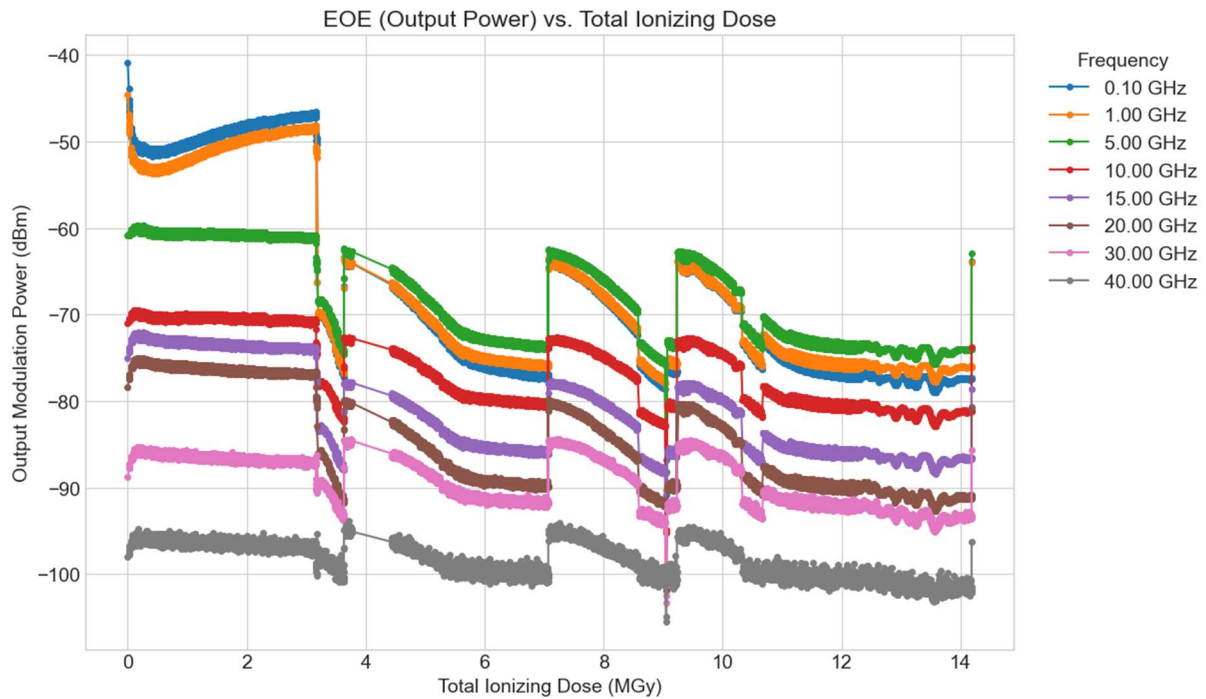


Figure 5.48: Electrical-optical-electrical response of the second shallow etched Mach-Zehnder modulator sample for different modulation frequencies and for an increasing total ionizing dose. Up until 3 MGy the modulator was accidentally driven in forward bias mode.

In this measurement initially the low frequencies drop quickly in their EOE response, but then gain some of that loss back. Higher frequencies just show a rise and then are mostly stagnant. At the changing point from FWB to RVB, the low frequencies show a drop of close to 20 dBm. For increasing frequencies this drop becomes progressively less pronounced. After that, the measurement behaves as the other measurement a radiation hardened MZM, except for windows, where the EOE response drops down. These are around 8.8 MGy and 10.5 MGy. After these windows, the EOE response recovers again.

Starting with the windows with the sudden drops: they are caused by a loss of the working point. The measurement was observed once when the drop in such a window happened. At the time the SA and other devices were hit by sunlight. The SA displayed a warning that it was measuring out of its calibrated range. However, this is not the behavior which is of interest in these curves. The interesting part is the range where the MZM was used with FWB modulation. After the initial swing the EOE response becomes very stable and for frequencies under 5 GHz the MZM delivers performance vastly better than for the RVB mode, which is highly unexpected. Due to time constraints, it could not be further investigated.

This is of note, since this opens the door for new applications within the detector framework. On the same chip as RMs or MZM, which aim to deliver very high data rates, a MZM modulating in FWB mode could be implemented for less data rate critical tasks or for chip health surveillance. The latter could for example exploit the relation between non-annealable and FWB modulation performance, since the device should be in an annealed state.

6 Summary and outlook

To take advantage of the future luminosity increases at the LHC at CERN, electronic readout systems must improve their data transmission rates. At the same time the increase in luminosity unavoidably corresponds to an increase in radiation and in consequence drives the demand for ever higher radiation hardness of those electronic systems. By attempting to fulfill these requirements the research naturally advances technologies both in and outside the field of high energy particle physics. The proposed solution for achieving this utilizes electro-optic modulators based on SOI technology, specifically implementing WDM. Each individual modulator encodes to one optical channel. (DE-)MUXs allow the aggregation of several channels onto one input and one output fiber for the whole system. This, accompanied by the general low space requirements enabled by the monolithic chips used, significantly reduces bulk. Here monolithic means the integration of driver electronics for the modulators and the modulators themselves on a single chip.

The irradiation measurements presented in this thesis represent a crucial step in the ongoing development of SiPh RMs and MZMs, especially targeted for use at the LHC. For the first time, in-situ measurements with regard to X-ray irradiation were conducted on a COTTONTAIL chip. The results conclusively prove that SiPh technology is a viable path forward when combined with FWB annealing. Specifically, the projected target maximum lifetime TID of 11 MGy was successfully verified for RMs at modulation frequencies of 100 MHz. For frequencies up to 40 GHz it could be shown that, if closer annealing intervals would have been chosen, that they most likely also would pass the 11 MGy threshold. Furthermore, the radiation hardened (shallow etched) MZMs demonstrated the same characteristics, suggesting they are robust candidates for high-TID environments. Additionally, it was proven that standard deep-etched MZMs are not viable for the use-case even when FWB annealing is deployed.

Adding to that, while no direct relation between TID and optimal working point shift (assuming its existence) could be proven, the possible range could be confined to the nanometer regime, possibly even sub-nanometer. With regard to the packaging and interconnection technology in combination with irradiation, the viability of flat fiber-to-chip coupling for heavily X-ray irradiated environments was proven, even with non-specialized fibers. Together with the permanent fixation of the fibers above the GCs via adhesive bonding, the measurements showed remarkable optical stability.

Continuing with the packaging and interconnection technology, but outside the irradiation framework several quantities of the COTTONTAIL and the flat fiber-to-chip coupling were characterized. These include the determination of the optimal angle for the facet angle-polishing of 41.17° , the formulation of a limit for the best-case optical coupling efficiency, by characterizing different waveguides with GCs at both ends, and the tuneability of the RMs via their integrated heaters.

Furthermore, certain aspects of the packaging and interconnection technology have been investigated and improved upon. For the electrical link between COTTONTAIL and measurement setups, wideband (DC to 40 GHz) working fan-out PCBs were designed and successfully deployed. Even with non-ideal wire bonds and non-RF, relatively low-cost PCB material, for all frequencies measurements were possible.

Finally, many new setups have been built, and new measurement software was written for them. Starting with a simple improvement to the chuck, which enables the angle-polishing of the fibers. By proposing an aluminum version, which will not degrade as fast as the previous solution made of plastic. Then, a fully new setup for the rotation alignment was built, enabling alignment of the fibers down to only 0.1° deviation from the ideal angle within the fiber holders. Also, the standard characterization setup was transferred and modified to work around the irradiation chamber. A full stack of code including drivers, data recording software, setup GUI and real time plots for automated measurements were created in Python.

Going from this, several future developments can be summarized. First to mention is the correction of code flaws, especially the working point control for the automated measurements and the integration of smart features for more precise tracking of the former. Then, longer measurements with shorter annealing intervals should be made to definitively prove that RMs and radiation hardened MZMs can pass the 11 MGy threshold at very high frequencies. Further, the handling process for gluing the fibers to the chip needs improvement. Currently it takes a large amount of time to produce a working sample. Generally, sample performance can be improved by optimizing the fan-out PCBs. For the MZMs specifically, one of the as-yet-unused alternative designs might offer better results. Also, an angle-optimization run needs to take place for the new fiber polishing chuck.

The focus in the near future lays on the integration of a so-called drop port into the RM layouts. A drop port is a second waveguide that catches some light, when the resonance condition is met and can be used for working point control. Additionally, future works towards a usable transmission system include drivers, which utilize the signal of a photodiode connected to the drop port, a modified fanout board or replacement suitable for the rest of the detector setup, and additional circuitry for biasing and annealing.

To summarize the content of this thesis: A photonic and electronic packaging methodology was developed or substantially improved to be able to build COTTONTAIL modulator samples, whose behavior under intense X-ray irradiation was investigated for the first time in-situ.

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Additional measurement data

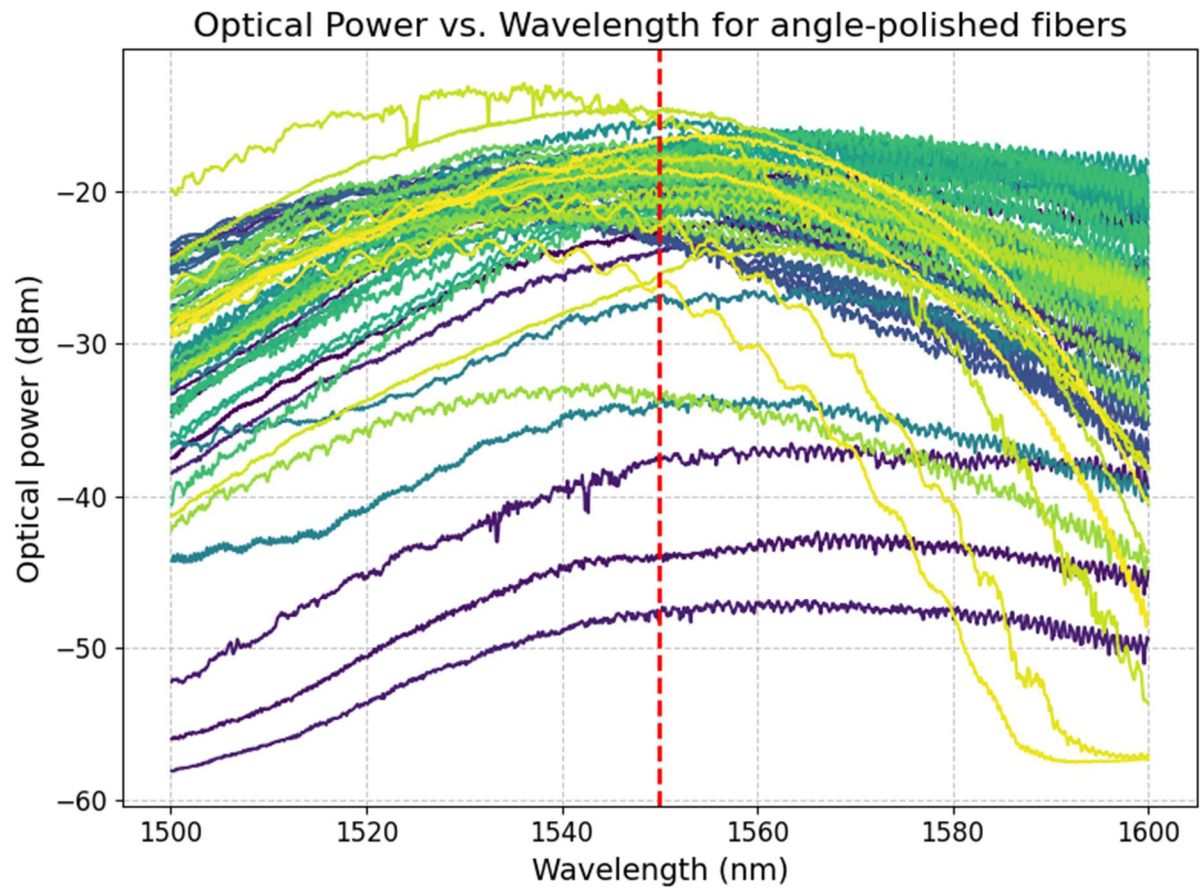


Figure 0.1: All measurements taken to optimize the chuck angle for the fiber polishing angle optimization done in section 4.2.

Set	Fiber type	Grinding/polishing date	Chuck angle	Theoretical fiber angle	Classification of max. pos.
1	Standard SMF	11.02.2025	42.27 °	8.00 °	Too high
2	Standard SMF	11.02.2025	42.27 °	8.00 °	Too high
3	Standard SMF	18.02.2025	41.27 °	11.23 °	Too low
4	Standard SMF	04.03.2025	42.00 °	8.79 °	-
5	Standard SMF	12.02.2025	42.00 °	8.79 °	Too high
6	PM	17.03.2025	42.00 °	8.79 °	In spec
7	PM	19.03.2025	42.00 °	8.79 °	In spec
8	PM	21.03.2025	42.00 °	8.79 °	Part. too high
9	PM	28.03.2025	42.00 °	8.79 °	Too high
10	PM	08.04.2025	41.17 °	11.23 °	Too low
11	PM	09.04.2025	41.17 °	11.23 °	In spec
12	Standard SMF	14.04.2025	41.17 °	11.23 °	-
13	Standard SMF	17.04.2025	41.17 °	11.23 °	-
14	Standard SMF	12.08.2025	41.17 °	11.23 °	-
			(part. metalized chuck)		
15	Standard SMF	13.08.2025	42.00 °	8.79 °	-
			(part. metalized chuck)		
16	Standard SMF	19.08.2025	41.17 °	11.23 °	-
17	Standard SMF	20.08.2025	42.00 °	8.79 °	-
18	PM	21.08.2025	42.00 °	8.79 °	-

Table 0.1: All angle-grinding/polishing parameters of the glass fiber sets, which produce the data shown in figure 7.1.

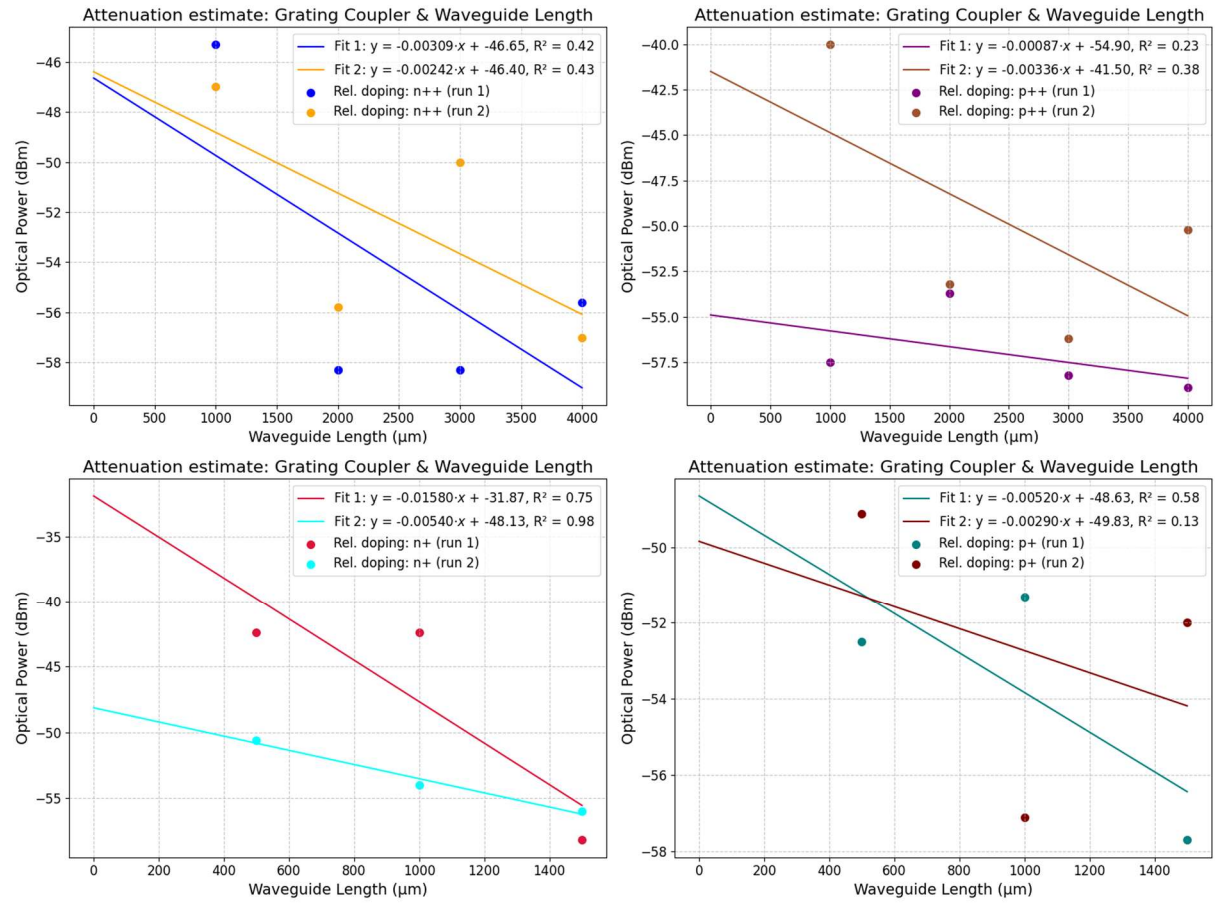


Figure 0.2: Analysis of the waveguide and grating coupler losses of the different relative doping concentrations as of Table 4.2.

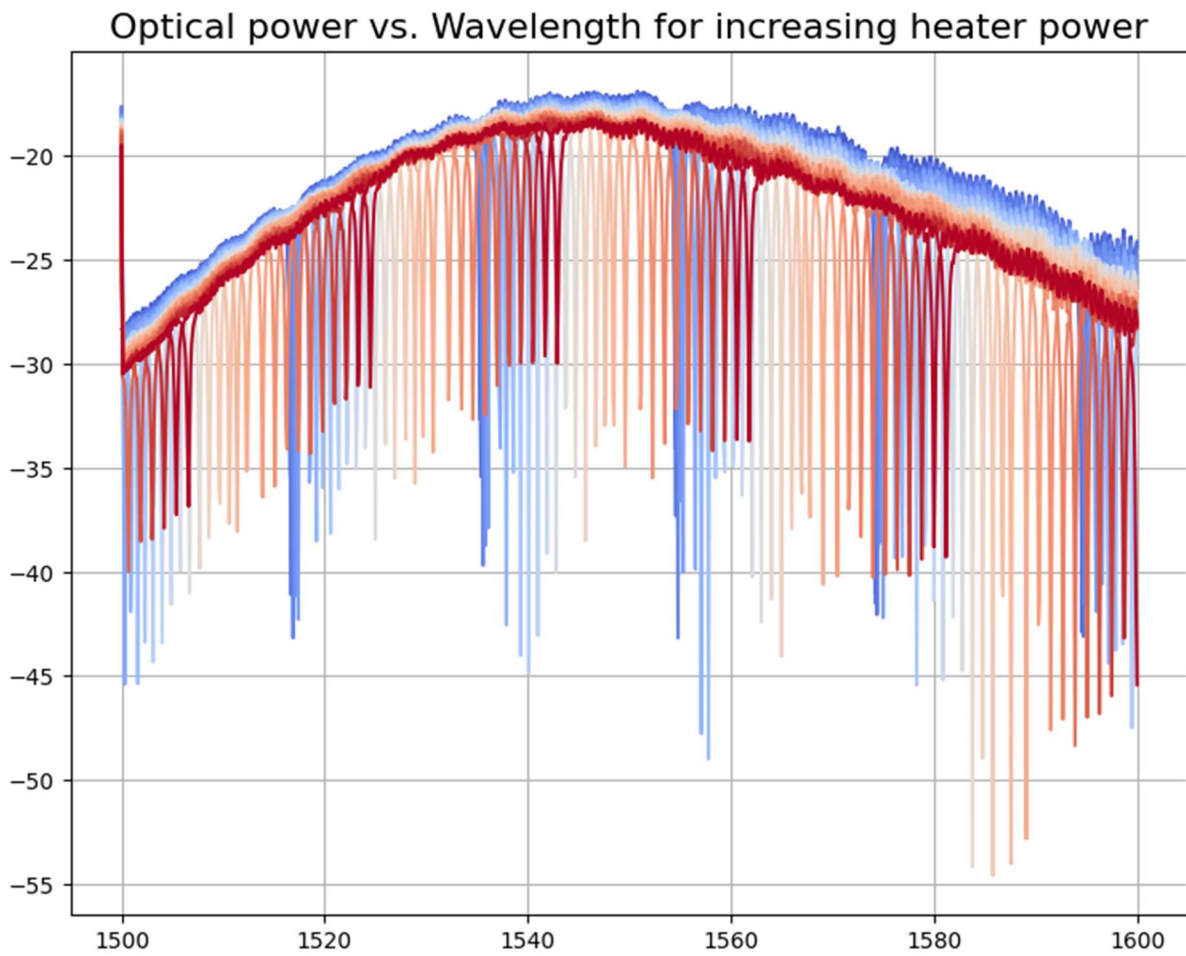


Figure 0.3: All recorded transmission curves of the heated ring modulator from section 4.3. Blue tones indicate a lower heating power (temperature) and red tones higher ones.

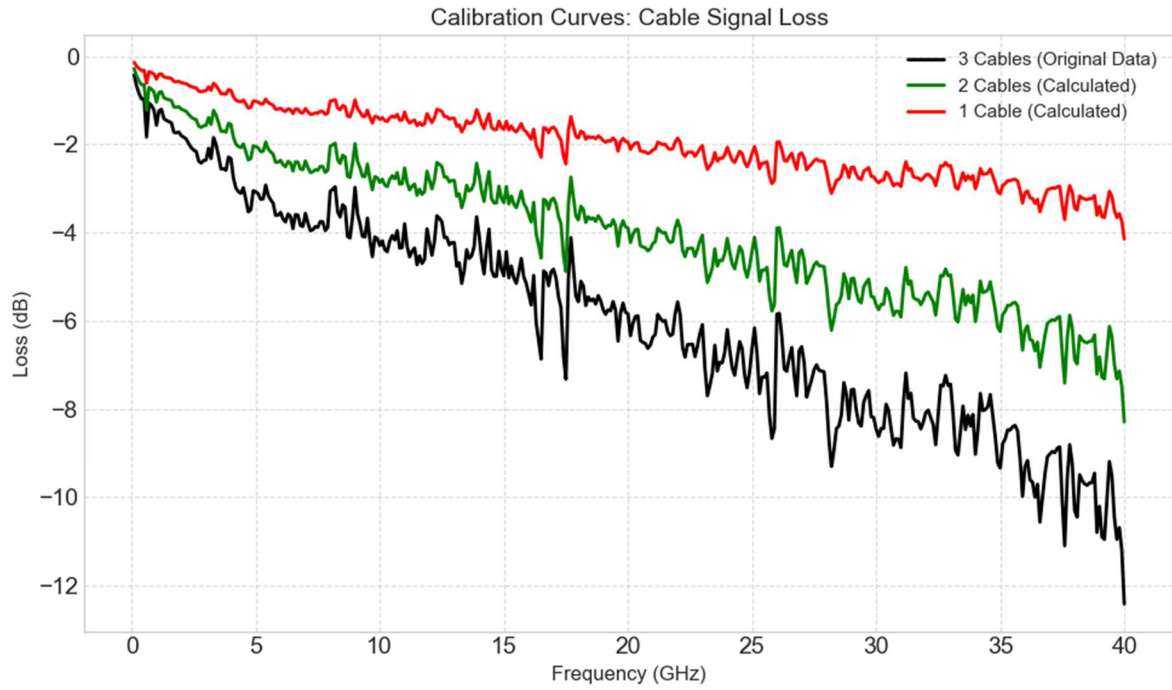


Figure 0.4: Calibration measurement to judge the attenuation of the used RF cables. The curve for three cables is the measurement result, while the others are derived mathematically (performed by M. Schneider (PhD)).

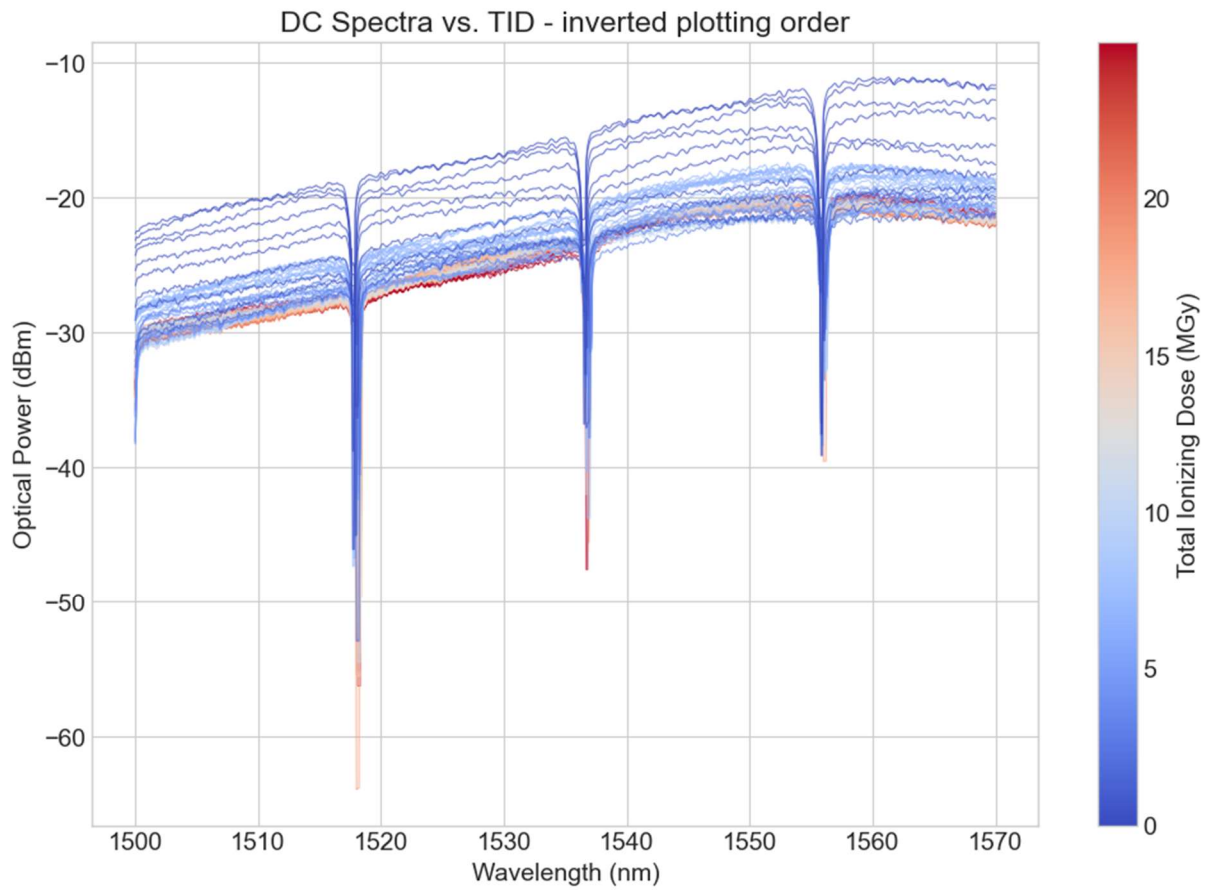


Figure 0.5: DC spectra evolution of the ring modulator sample for an increasing total ionizing dose. Lower total ionization doses are printed after high ones, so that the “swings” can be seen better.

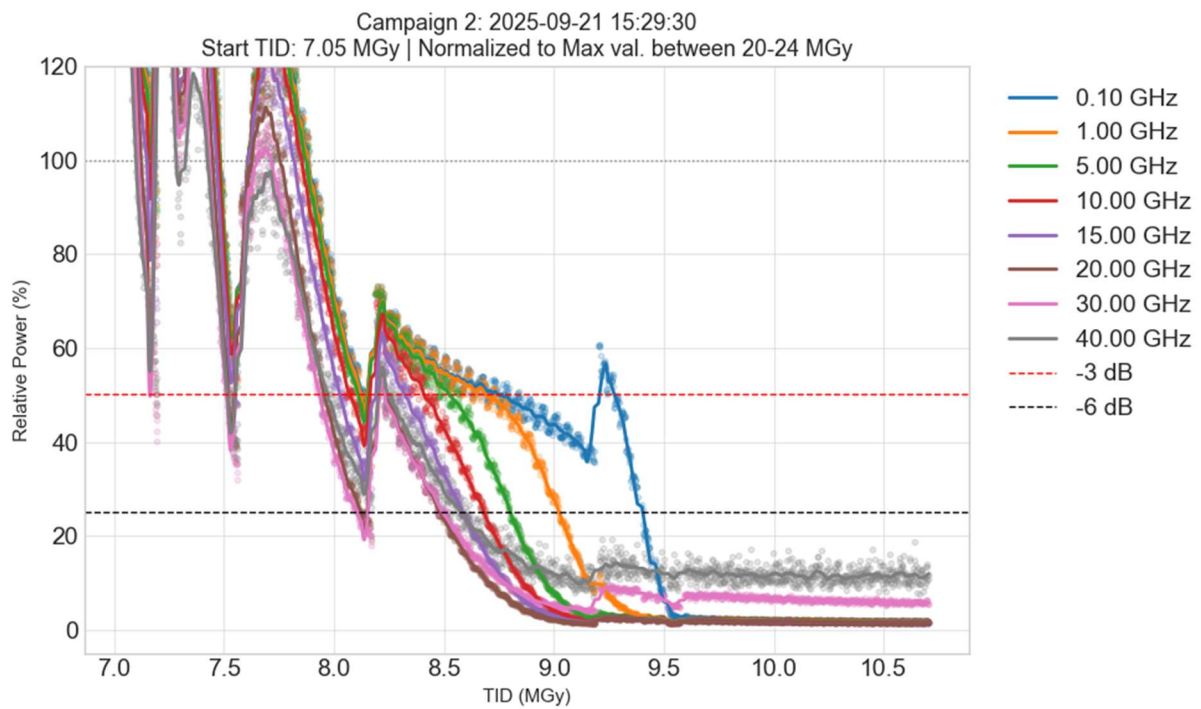


Figure 0.6: The second irradiation campaign of the ring modulator sample.

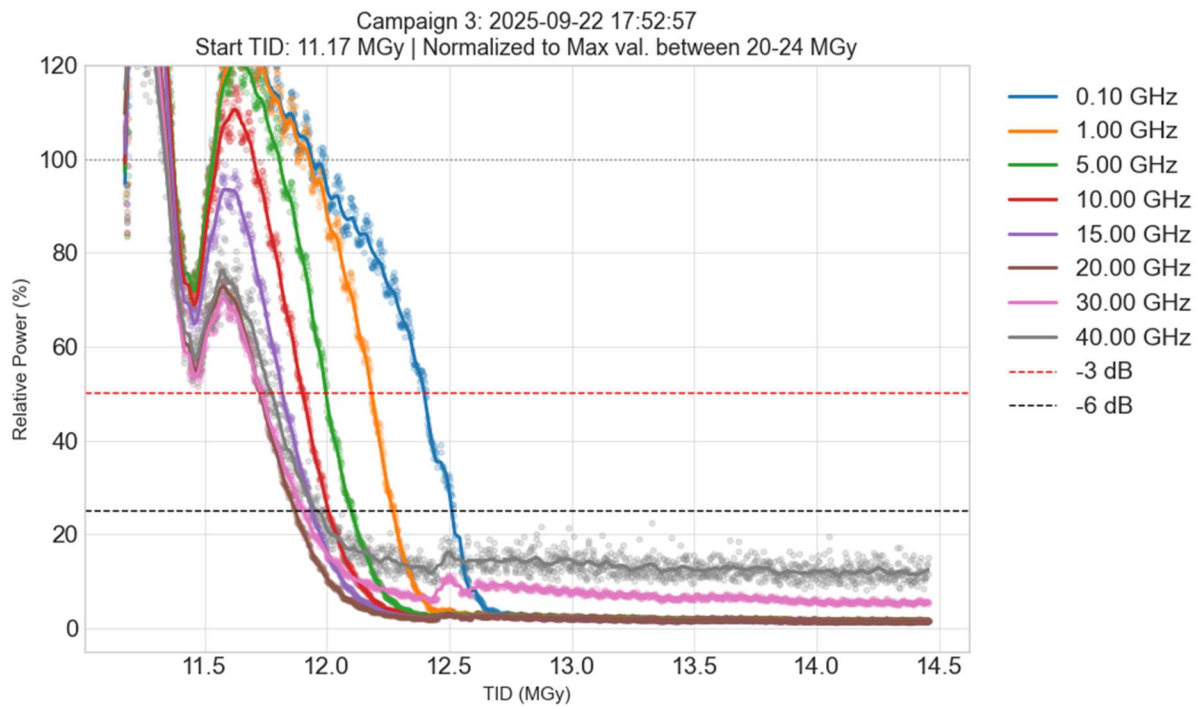


Figure 0.7: The third irradiation campaign of the ring modulator sample.

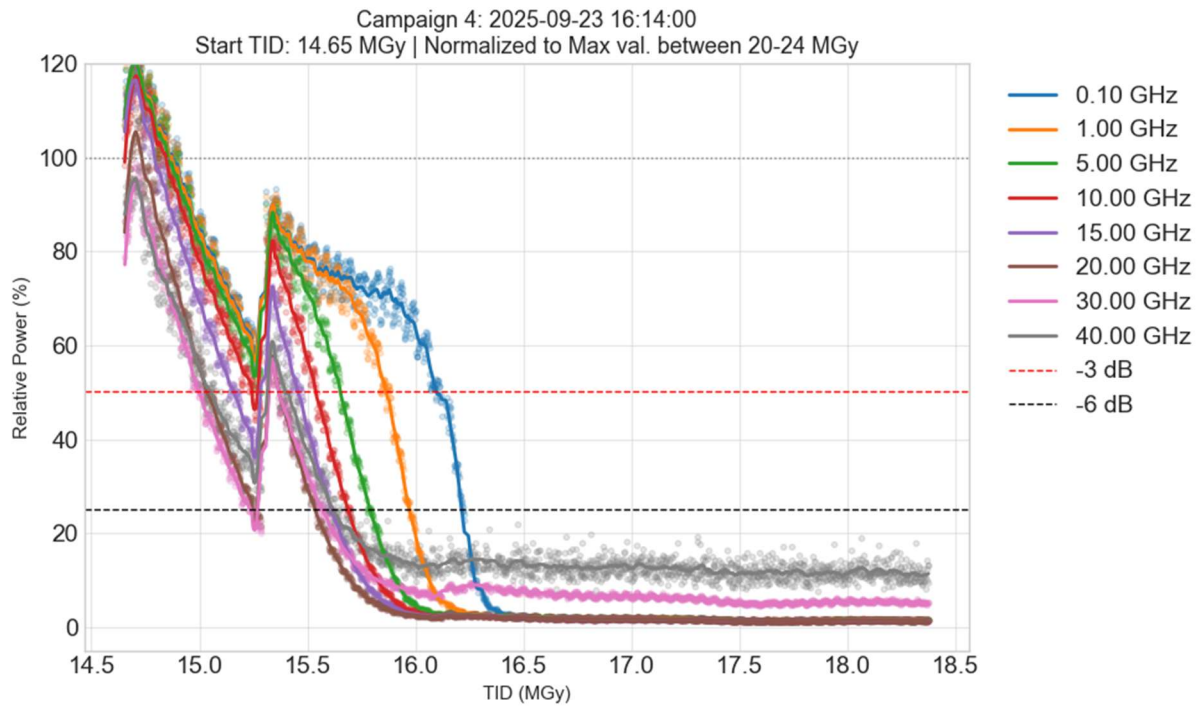


Figure 0.8: The fourth irradiation campaign of the ring modulator sample.

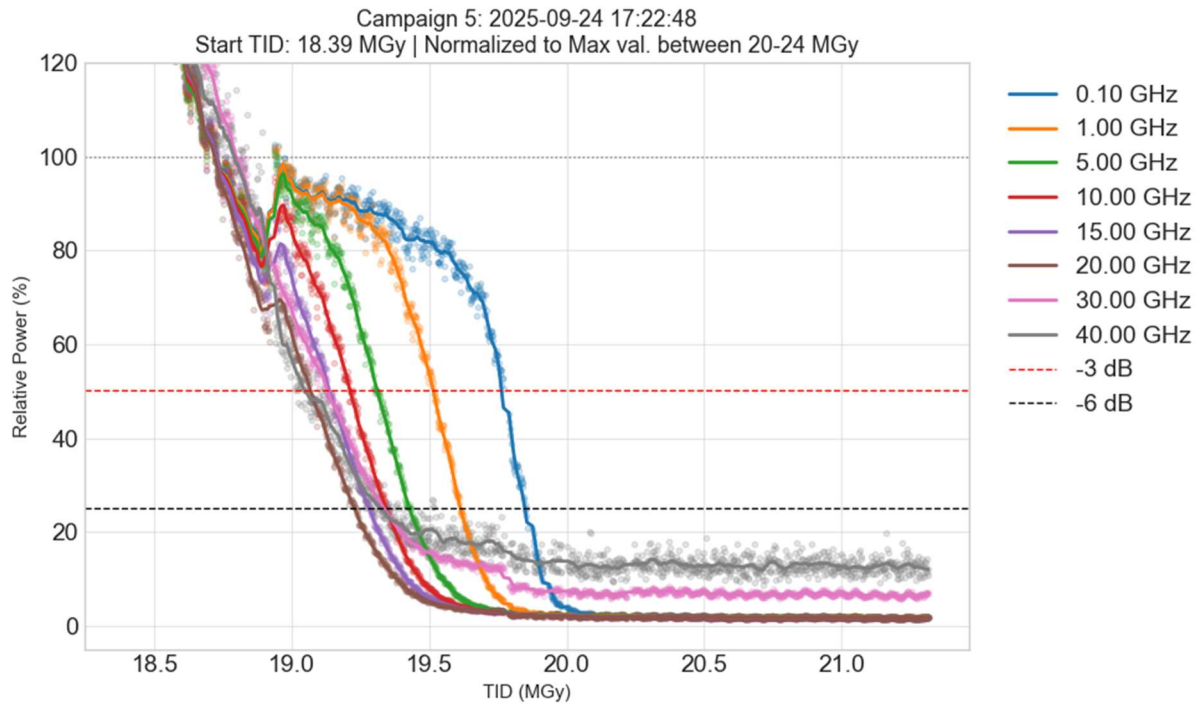


Figure 0.9: The fifth irradiation campaign of the ring modulator sample.

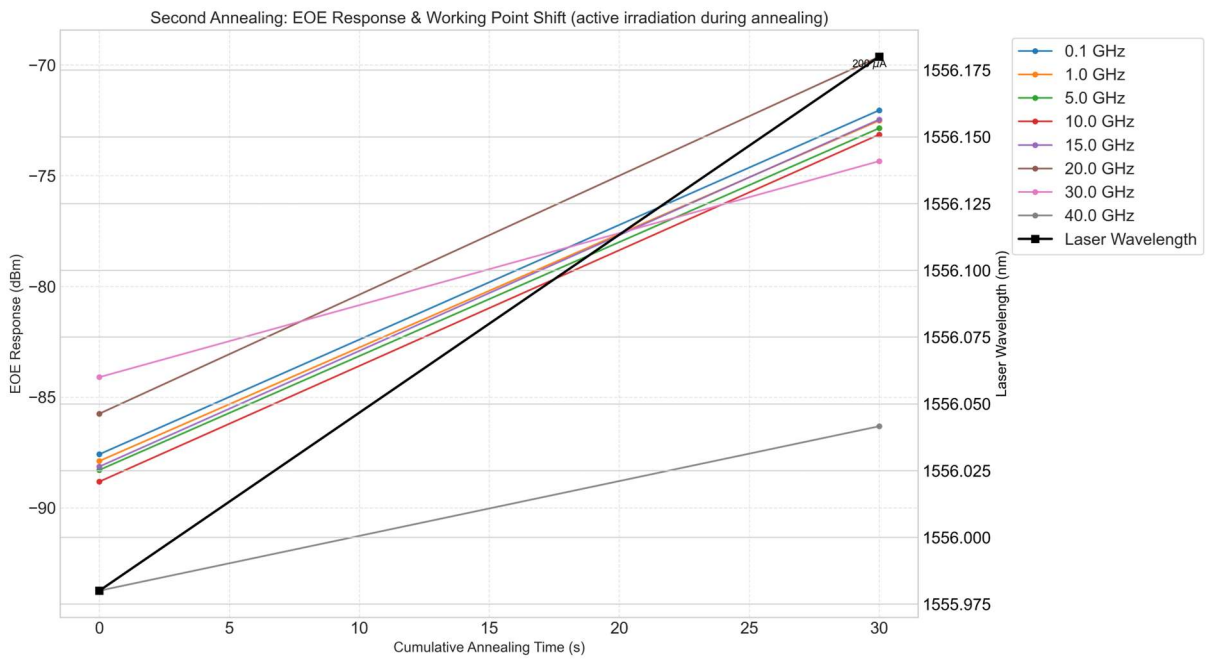


Figure 0.10: Second annealing process of the ring modulator sample. The single annealing step was done with 200 μ A.

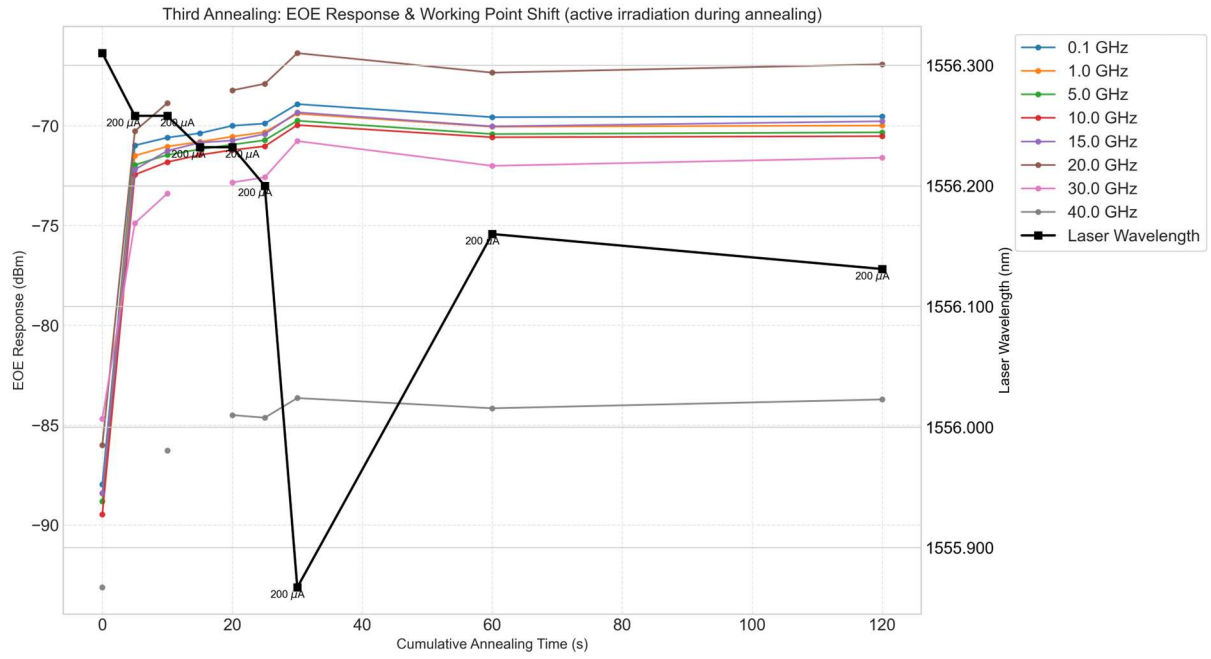


Figure 0.11: Third annealing process of the ring modulator sample. All annealing steps were done with $200\ \mu\text{A}$.

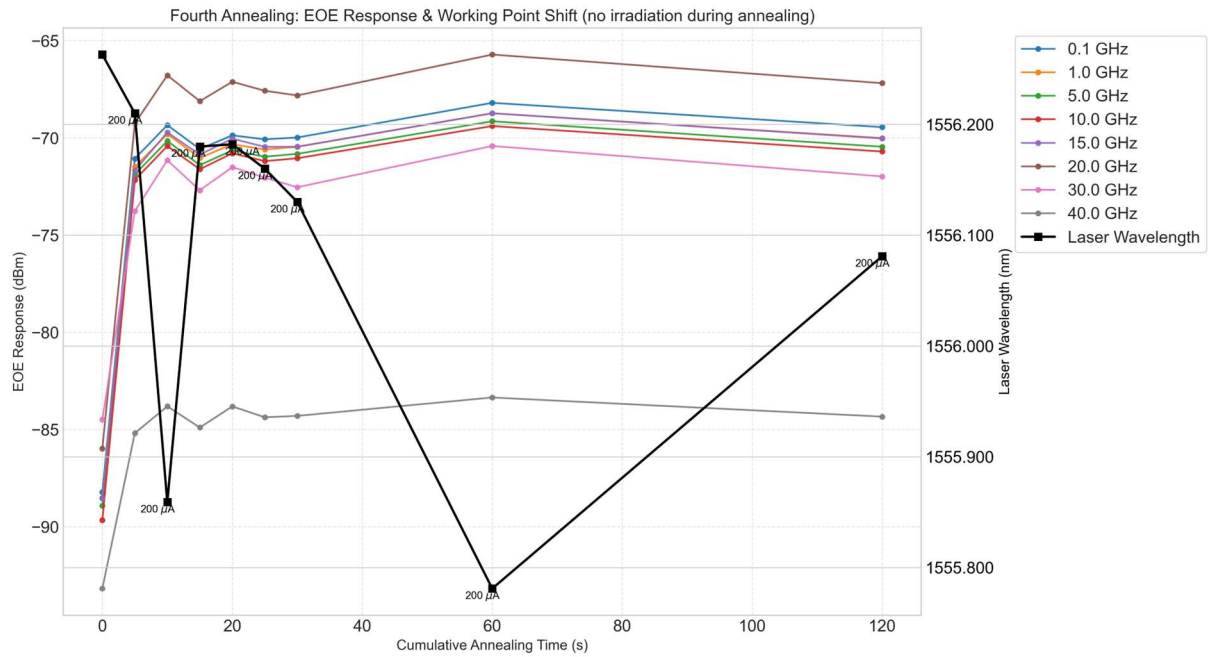


Figure 0.12: Fourth annealing process of the ring modulator sample. All annealing steps were done with $200\ \mu\text{A}$.

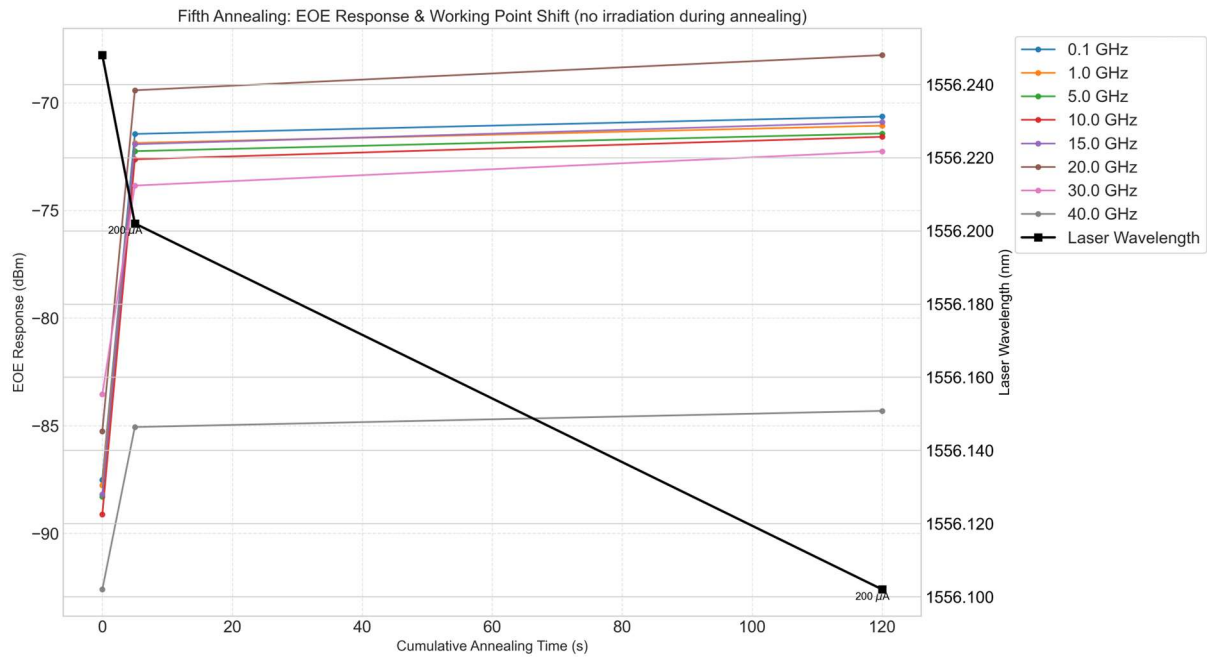


Figure 0.13: Fifth annealing process of the ring modulator sample. All annealing steps were done with 200 μ A.

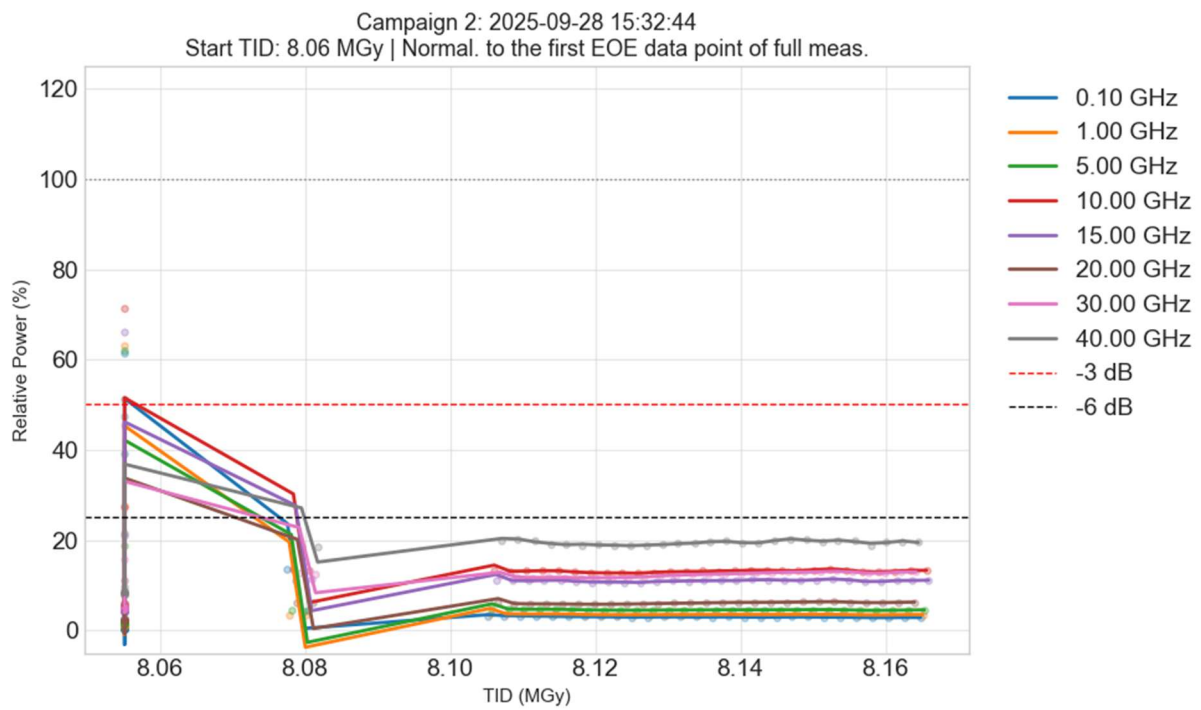


Figure 0.14: The second irradiation campaign of the deep etched Mach-Zehnder modulator sample.

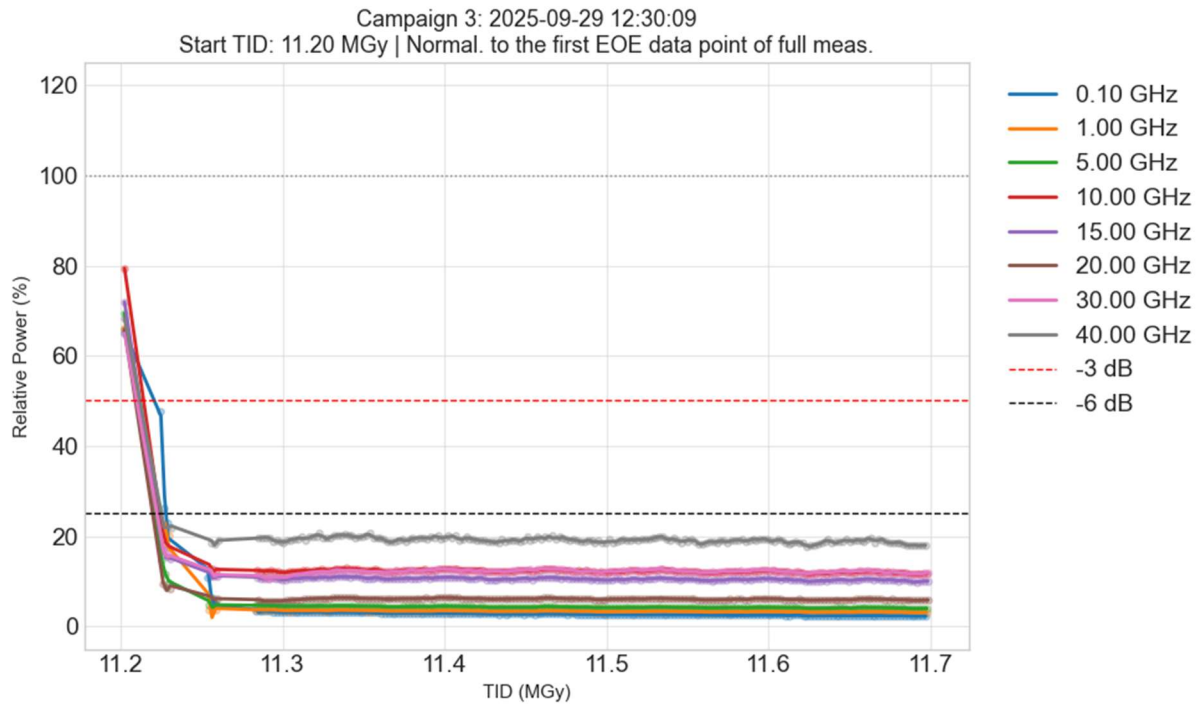


Figure 0.15: The third irradiation campaign of the deep etched Mach-Zehnder modulator sample.

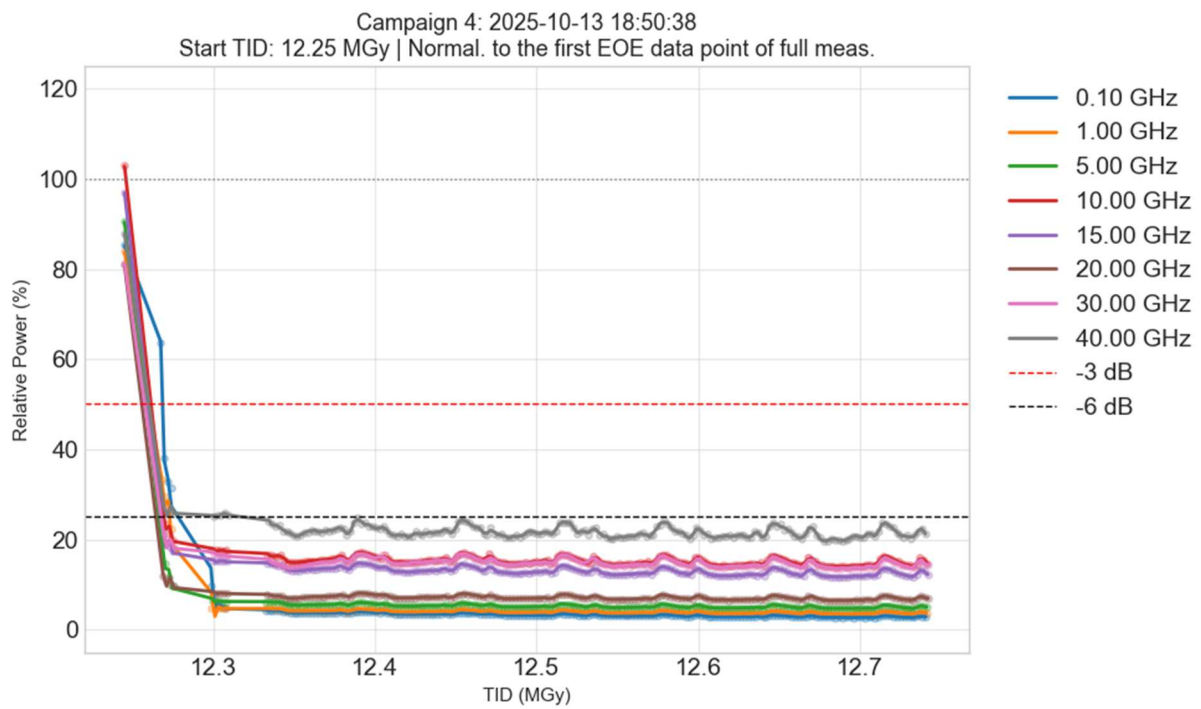


Figure 0.16: The fourth irradiation campaign of the deep etched Mach-Zehnder modulator sample.

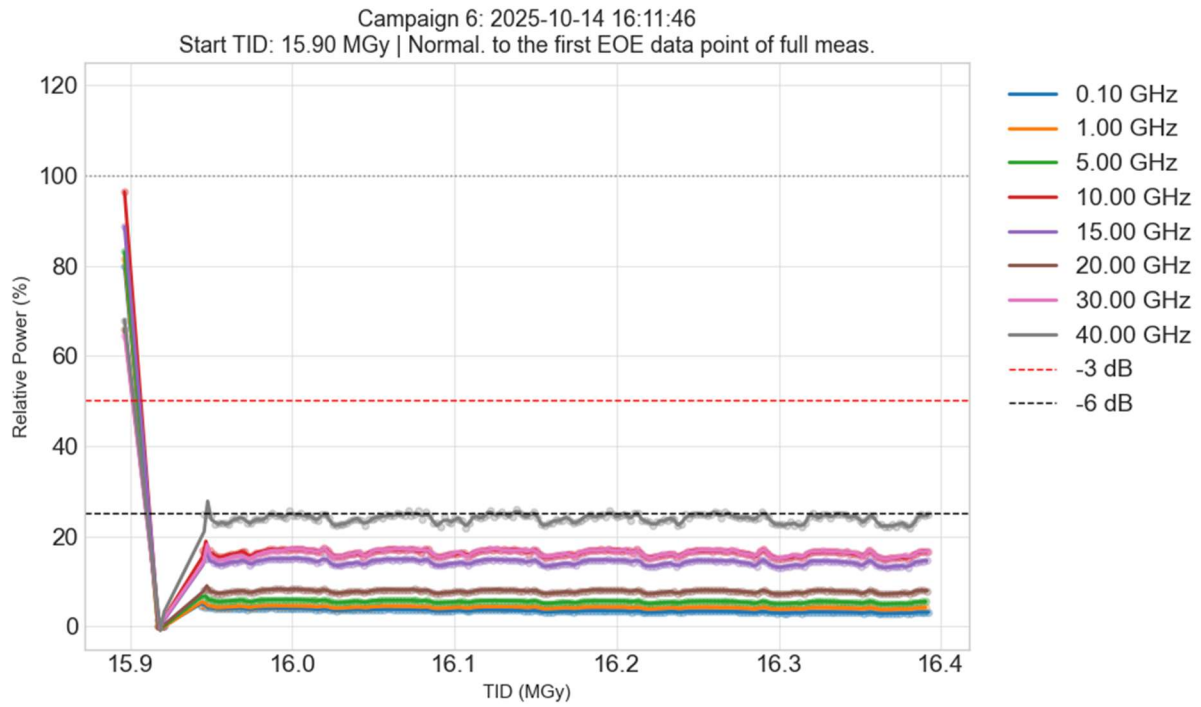


Figure 0.17: The sixth (final) irradiation campaign of the deep etched Mach-Zehnder modulator sample.

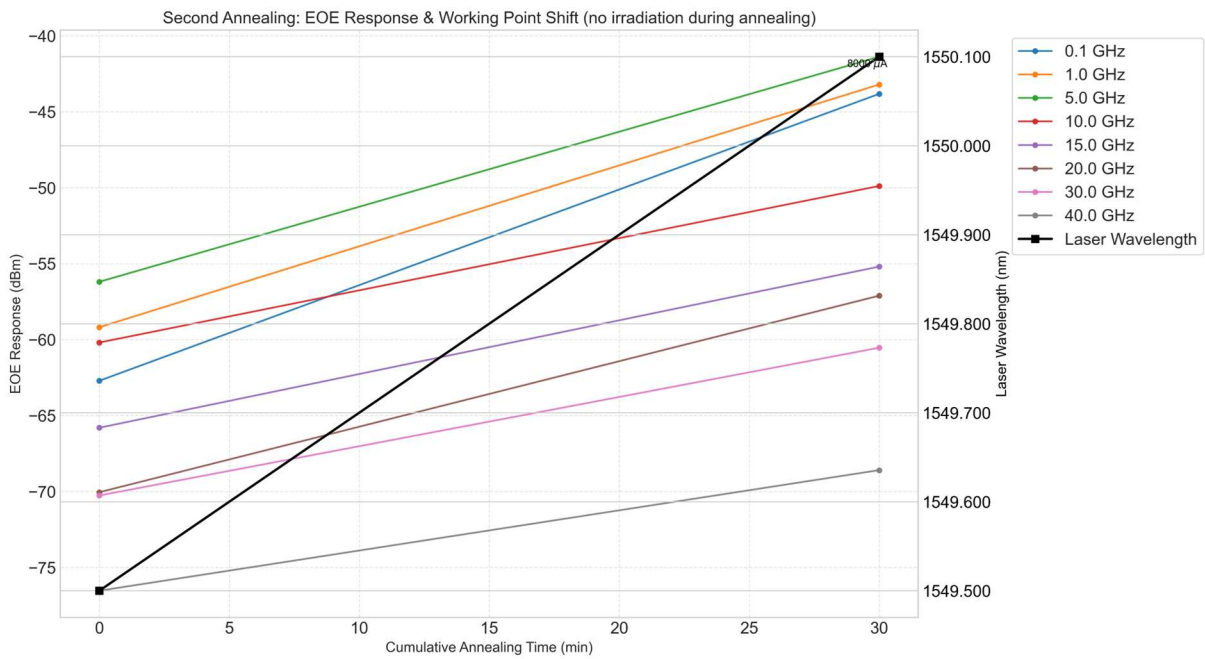


Figure 0.18: Second annealing process of the deep etched Mach-Zehnder modulator sample. The single annealing step was done with 8000 μA .

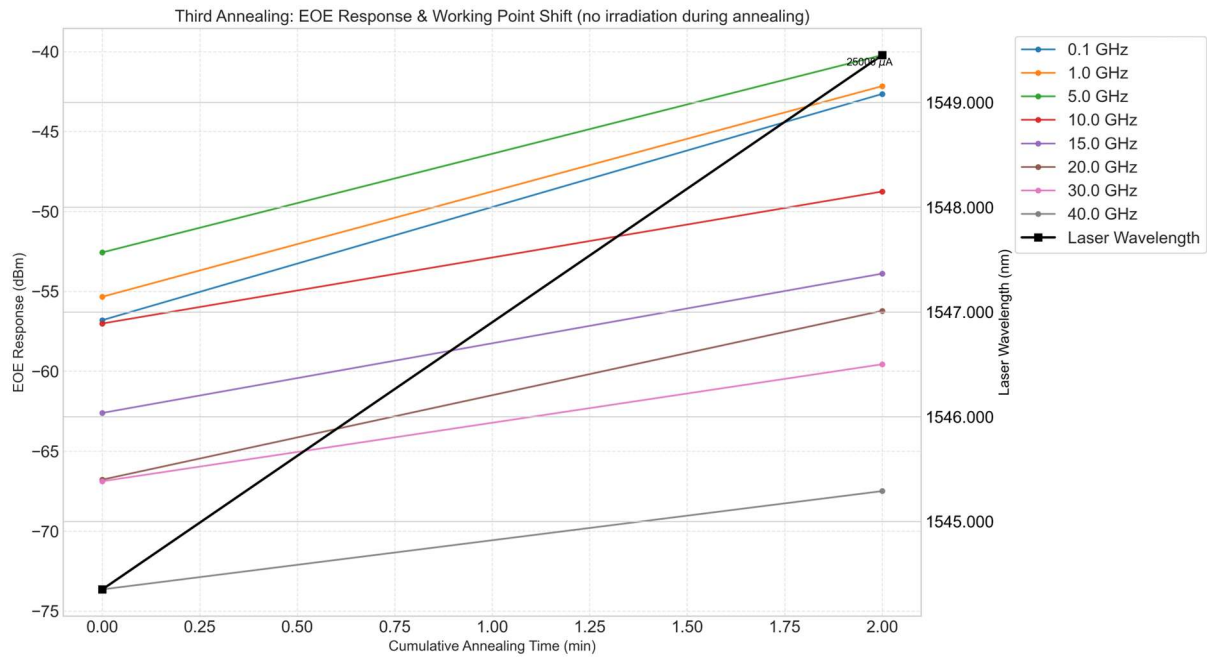


Figure 0.19: Third annealing process of the deep etched Mach-Zehnder modulator sample. The single annealing step was done with 25000 μA .

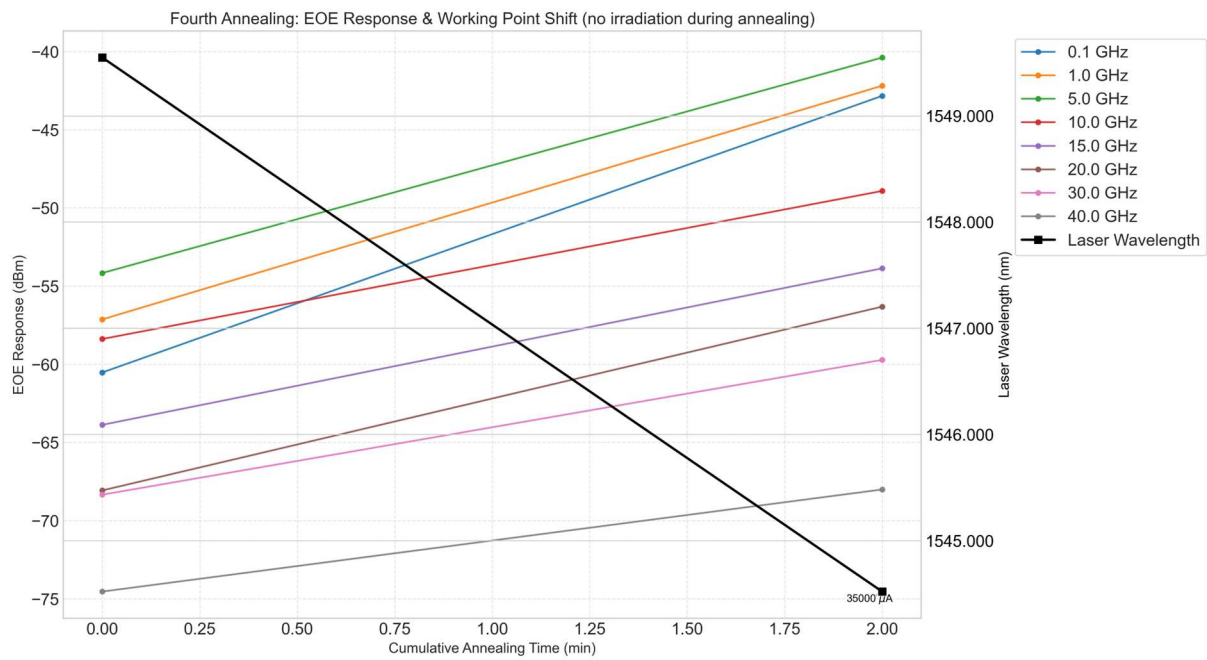


Figure 0.20: Fourth annealing process of the deep etched Mach-Zehnder modulator sample. The single annealing step was done with 35000 μA .

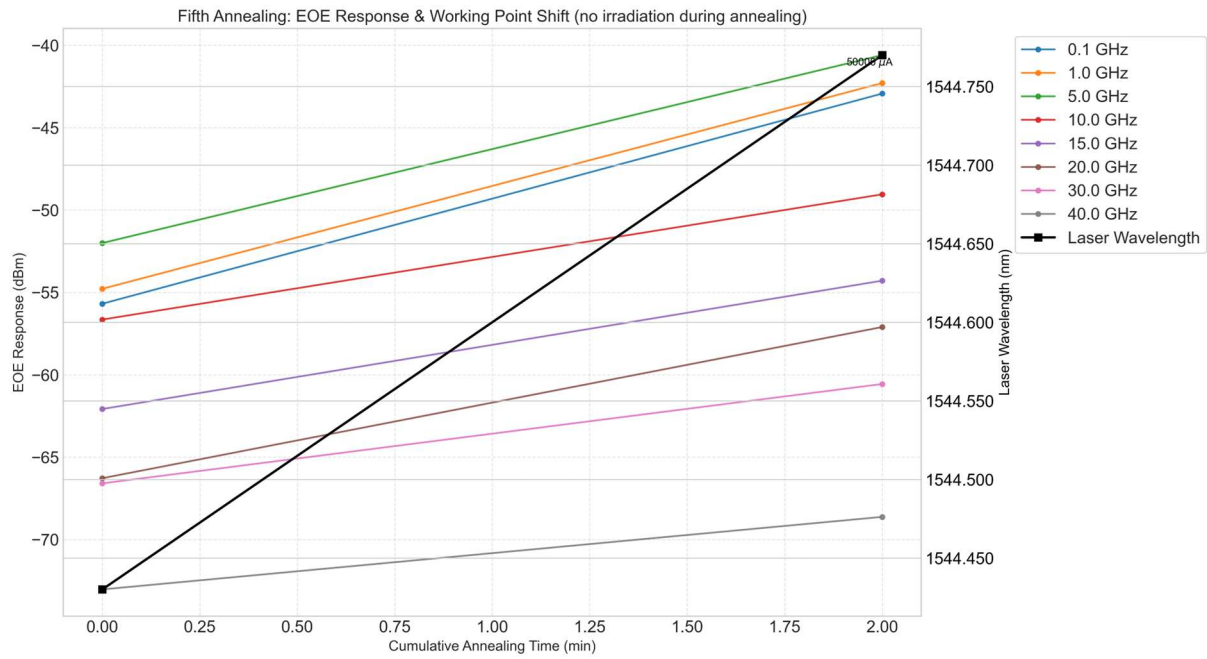


Figure 0.21: Final (fifth) annealing process of the deep etched Mach-Zehnder modulator sample. The single annealing step was done with 50000 μA .

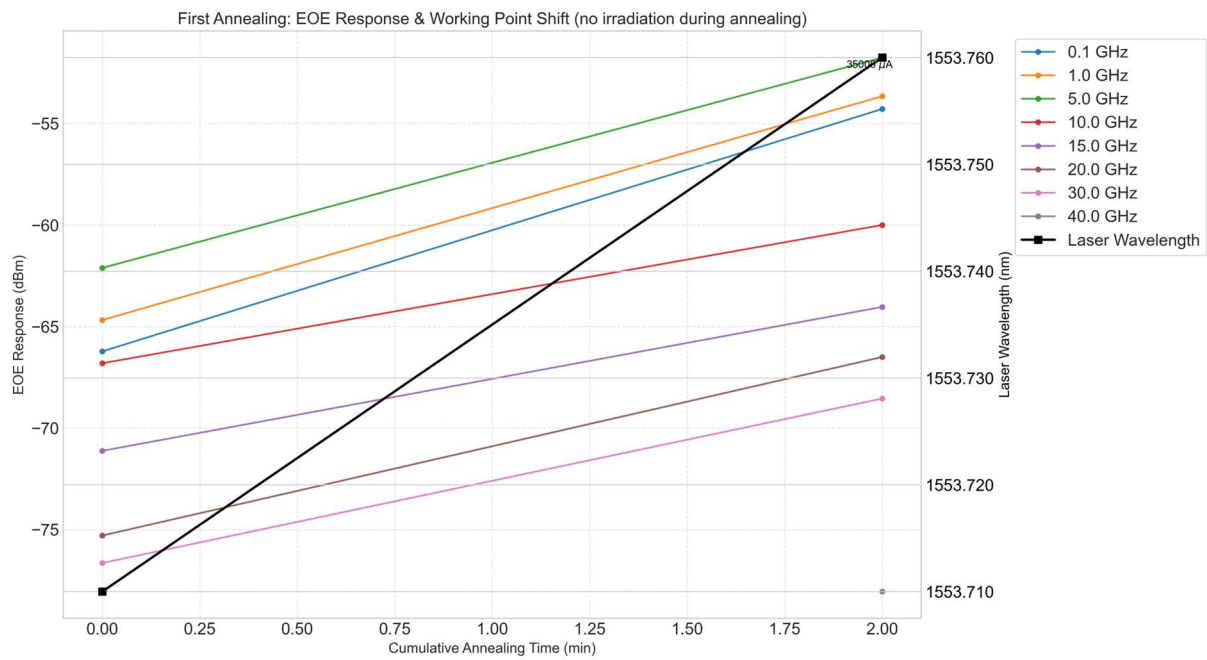


Figure 0.22: First annealing process of the first shallow etched Mach-Zehnder modulator sample. The single annealing step was done with 35000 μA .

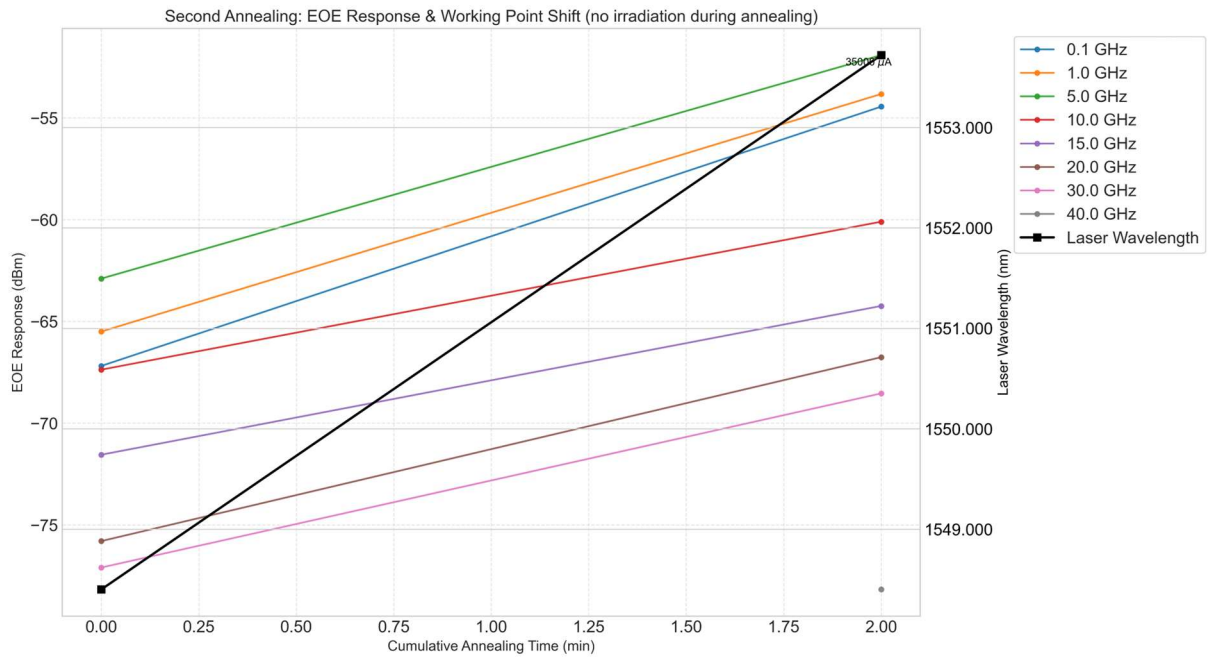


Figure 0.23: Final (second) annealing process of the first shallow etched Mach-Zehnder modulator sample. The single annealing step was done with 35000 μA .

Supporting materials

Clarifications on the derivation of optical modes in a waveguide:

There are two approaches which both lead to the same solution. One method looks at the phase differences between a full roundtrip (reflection at the first and second medium) [13]. We can conclude that in the y-axis we need a phase difference of $m \cdot 2\pi$, where m is an integer, since any other waves would cancel each other out during propagation. This leads to

$$2k_y h - 2\phi_r = 2m\pi, \quad (0.1)$$

where k_y is the propagation constant in y-direction and ϕ_r an additional phase from the reflection. The latter stems from the fact that the reflection does not occur directly at the interface. This corresponds to the field penetrating the interface in the other method below. So, we have only discrete allowed modes within our waveguide.

For the other method we need to consider a transverse electric (TE) wave like

$$E = E_0 e^{i(k_z z \pm \omega t)}. \quad (0.2)$$

Additionally, we fix the electric field to the xz-plane. The electric field of this wave doesn't have any component in the propagation direction, meaning along the z-axis in our example. Using the wave equation for the electric field

$$\nabla^2 E = \mu_m \epsilon_m \frac{\partial^2 E}{\partial t^2} \quad (0.3)$$

after some calculations [13] we arrive at

$$\frac{\partial^2 E_x}{\partial y^2} = (k_z^2 - k_0^2 n_i^2) E_x. \quad (0.4)$$

This equation is valid for the electric field in each medium and describes the electric field along the x-axis. k_0 is the propagation constant in free space. Next, we form three independent equations with n_1 , n_2 and n_3 . To solve this problem, we use boundary conditions. These are that both the electric field and the first derivative to y are continuous at the media interfaces. The latter is a requirement from the Maxwell equation

$$\nabla \times E = -\frac{\partial B}{\partial t} \quad (0.5)$$

and ensures the continuity of the magnetic field. Doing the necessary calculations [52], we arrive at

$$\kappa \tan\left(\kappa \frac{h}{2}\right) = \gamma \quad \text{and} \quad \kappa \cot\left(\kappa \frac{h}{2}\right) = -\gamma. \quad (0.6)$$

The first equation is for symmetric modes and the second one for antisymmetric ones. $\kappa = k_y$ is the transverse phase constant in the core and γ the transverse decay constant in the cladding materials. The mode count m is here given by how many independent solutions both equations can yield.

A final step would be the calculation of the effective refractive index n_{eff} of each mode by calculating k_z . First, an expression excluding k_z for either γ or κ must be found. In the next step the corresponding equation in (7.6) needs to be solved, usually numerically or graphically. With the numerical value of γ or κ we can then get k_z . Then effective refractive n_{eff} index can then be formulated as (2.6).

Overview over the unused Mach-Zehnder fan-out connection (bias-tee) PCBs:

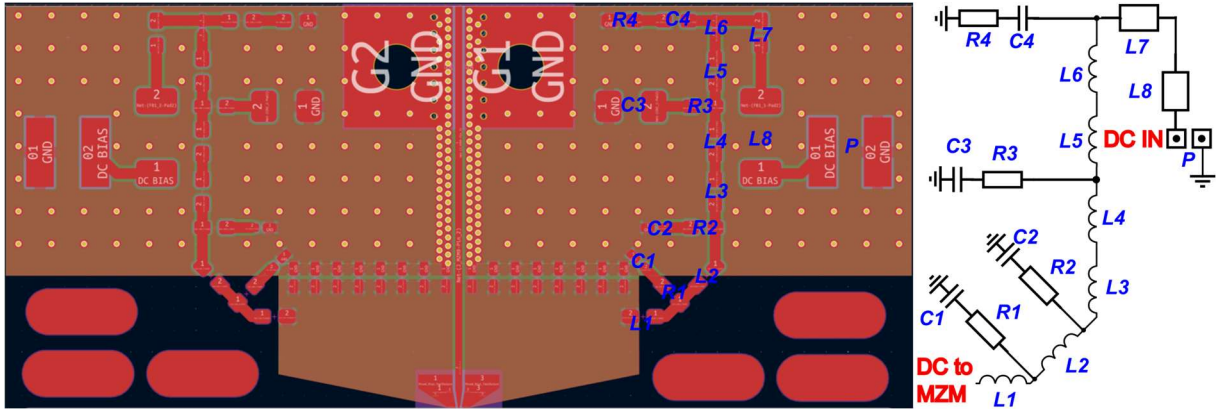


Figure 0.24: A bias-tee DC arm layout with many fine-tuned steps.

Designation in figure 7.24	Component type	Component name
L8	3.2 μH inductor (ferrite core)	Laird HZ1206C202R-10
L7	350 nH inductor (ferrite core)	Würth 742843122
L6	1.0 μH inductor	Murata LQW15CA1R0K00D
L5	100 nH inductor	Murata LQW15AWR10G80D
L4	12 nH inductor	Würth 7447650120
L3	2.2 nH inductor	Kyocera AVX L04022R2BHNTR
L2	0.82 nH inductor	Kyocera AVXL0402R82BHNTR
L1	0.2 nH inductor	Murata LQP02HQ0N2B02E
R4	0 Ω jumper	Panasonic ERJ-2GE0R00X
R3	340 Ω resistor	Panasonic ERA-2AEB3400X
R2	75 Ω resistor	Vishay CH0402-75RGFPT
R1	200 Ω resistor	Vishay CH0402-200RGFPA
C4	10 nF capacitor	TDK C1005X7S2A103K050BB
C3	100 pF capacitor	Walsin 0805N101FW251X
C2	0.5 pF capacitor	Kyocera AVX 02011J0R5ZBSTR
C1	0.05 pF capacitor	Kyocera AVX 02011JR05ZBSTR
P	Connection pins	Samtec TSM-102-01-T-SH

The goal of this design was to have a chain of inductors with increasingly higher SRFs so that no parasitics could get through to the Mach-Zehnder modulator. The R-C side arms should effectively de-Q any parasitics that may have passed the previous inductor step(s).

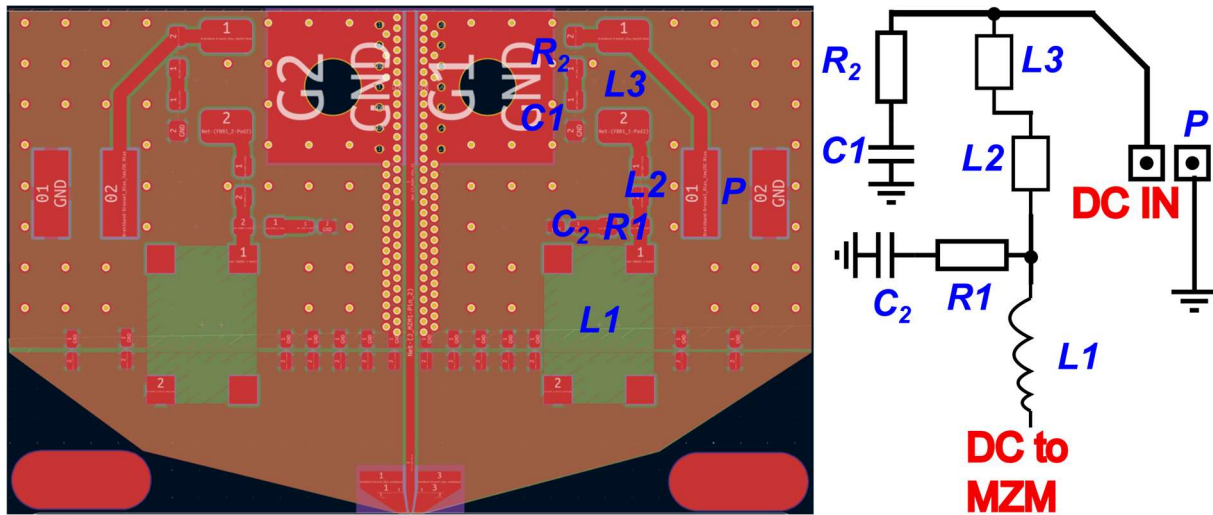


Figure 0.25: A bias-tee DC arm layout with a SMD broadband conical inductor.

Designation in figure 7.25	Component type	Component name
L3	3.2 μ H inductor (ferrite core)	Laird HZ1206C202R-10
L2	350 nH inductor (ferrite core)	Würth 742843122
L1	530 nH broadband conical ind.	Coilcraft BCR-531JLC
R2	0 Ω jumper	Panasonic ERJ-2GE0R00X
R1	75 Ω resistor	Vishay CH0402-75RGFPT
C2	10 nF capacitor	TDK C1005X7S2A103K050BB
C1	0.5 pF capacitor	Kyocera AVX 02011J0R5ZBSTR
P	Connection pins	Samtec TSM-102-01-T-SH

Broadband conical inductors often cover a 1 GHz to 40 GHz range (in terms of SRF). This design offers a footprint for such SMD-housed conical inductors. Only the lower frequency range needs to be covered by additional components.

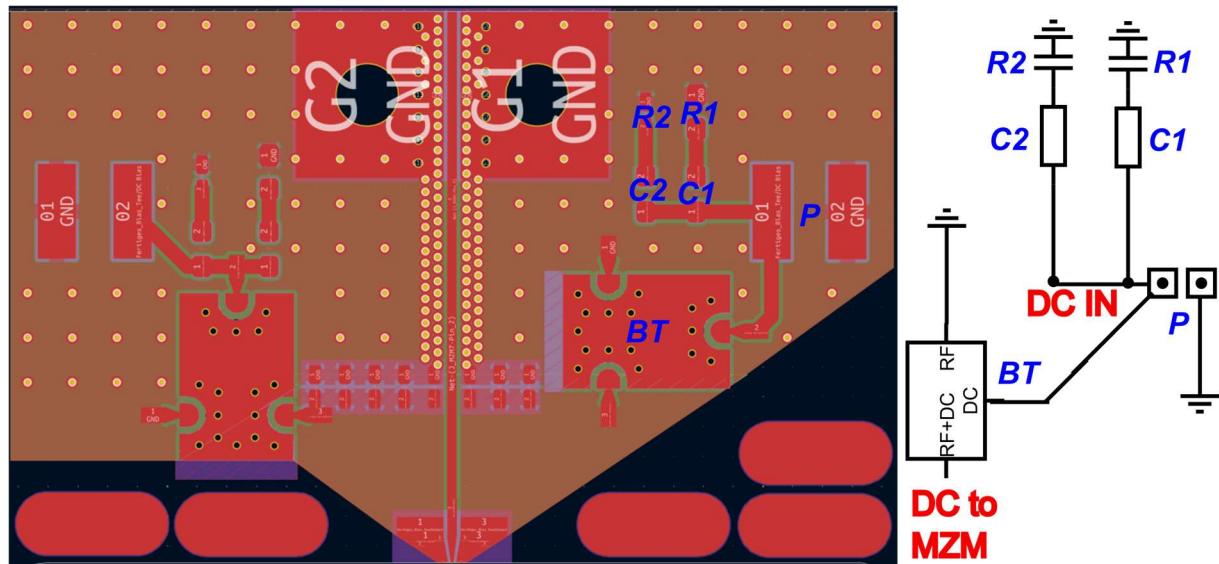


Figure 0.26: A bias-tee DC arm layout using a bias-tee IC.

Designation in figure 7.26	Component type	Component name
BT	Integrated bias-tee	Marki Microwave BT-0014SMG-2
R2	0 Ω jumper	Panasonic ERJ-2GE0R00X
R1	75 Ω resistor	Vishay CH0402-75RGFPT
C2	10 nF capacitor	TDK C1005X7S2A103K050BB
C1	0.5 pF capacitor	Kyocera AVX 02011J0R5ZBSTR
P	Connection pins	Samtec TSM-102-01-T-SH

Integrated bias-tee chips exist and were available at the institute. If it had been the case that the self-made designs would not work, this was created as an alternative approach.

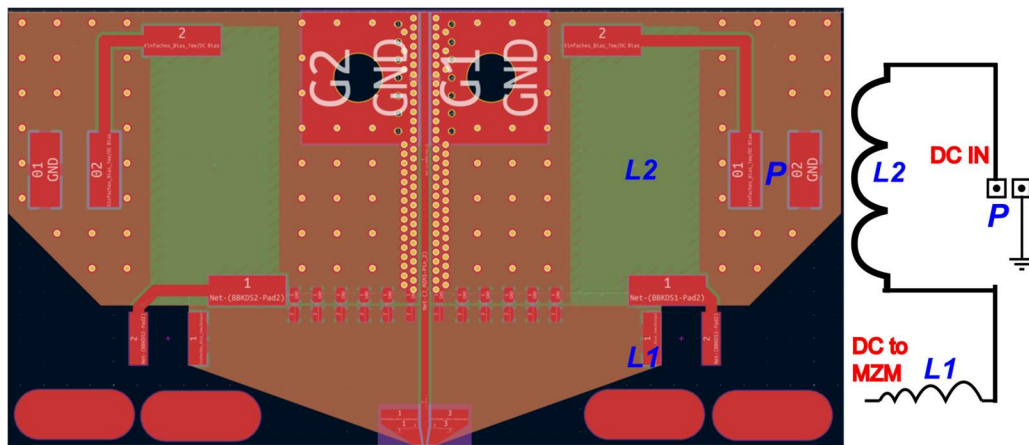


Figure 0.27: A simple bias-tee layout using a broadband conical inductor.

Designation in figure 7.27	Component type	Component name
L2	3.5 μ H inductor	e.g. Coilcraft 4310LC-352KEC
L1	530 nH broadband conical ind.	e.g. Coilcraft BCL-531JL
P	Connection pins	Samtec TSM-102-01-T-SH

Many non-SMD bias-tees are very simple. They use a normal, big inductor coil followed by a broadband conical inductor. The footprint for the conical inductor was design for unboxed conical inductors but may also be used with SMD-housed ones.

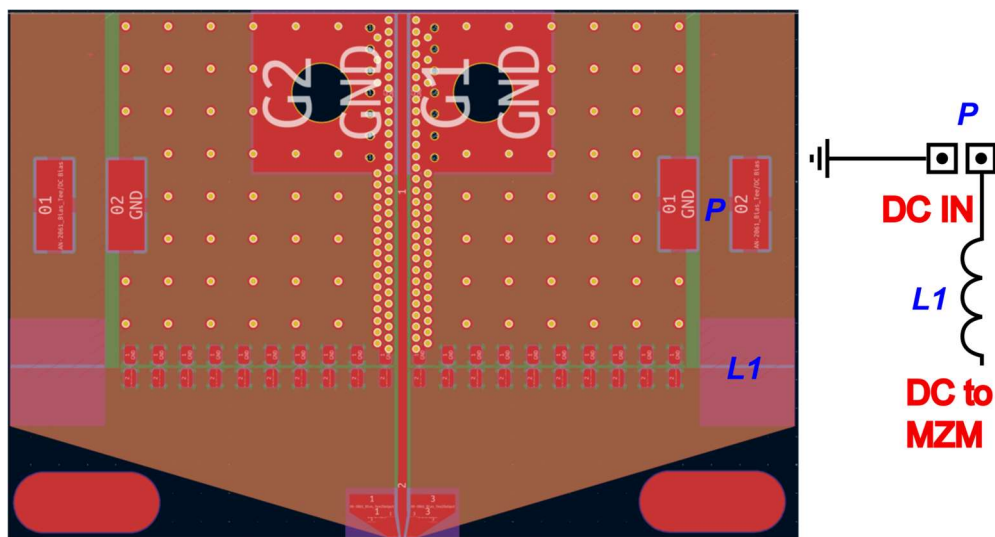


Figure 0.28: A testing layout.

Designation in figure 7.28	Component type	Component name
L1	Any inductor	None
P	Connection pins	Samtec TSM-102-01-T-SH

A testing platform

