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Masterthesis

Feasibility Study for High Power RF – Energy Recovery in Particle Accelerators

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Abstract

When dealing with particle accelerators, especially in systems with travelling wave structures and low beam loading, a substantial amount of RF power is dissipated in 50Ω termination loads. For the Super Proton Synchrotron (SPS) at Cern this is 69 % of the incident RF power or about 1 MW. Different ideas, making use of that otherwise dissipated power, are presented and their feasibility is reviewed. The most feasible one, utilizing an array of semiconductor based RF/DC modules, is used to create a design concept for energy recovery in the SPS. The modules are required to operate at high power, high efficiency and with low harmonic radiation. Besides the actual RF rectifier, they contain additional components to ensure a graceful degradation of the overall system. Different rectifier architectures and semiconductor devices are compared and the most suitable ones are chosen. Two prototype devices were built and operated with up to 400 W of pulsed RF power. Broadband measurements – capturing all harmonics up to 1 GHz – were done in the time domain with a fast sampling oscilloscope. The data was processed by a Matlab script, which allowed to extract the efficiency, the reflected power spectrum and the complex reflection coefficient, all dependent on input power level. Systematic errors were compensated by an open-, short-, load- calibration and the reflection coefficient was plotted in a smith chart, making the measurement setup equivalent to a single port, large signal vector network analyzer, operating with high power RF pulses. The RF/DC converter prototype with two Schottky diodes reached 88.7% efficiency at an input power level of 284 W.

Eidesstattliche Erklärung

Hiermit versichere ich, die vorliegende Masterthesis ohne unzulässige fremde Hilfe selbständig verfasst und keine anderen als die angegebenen Quellen und Hilfsmittel benutzt zu haben.

Genf, den 24.09.10

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Engineering is not merely knowing and being knowledgeable, like a walking encyclopedia; engineering is not merely analysis; engineering is not merely the possession of the capacity to get elegant solutions to non-existent engineering problems; engineering is practicing the art of the organized forcing of technological change... Engineers operate at the interface between science and society...

Gordon Stanley Brown

1. Introduction

1.1 Cern

At a time, just after the second world war, nuclear physics in Europe was in a very bad state. In this period, a handful of scientists – amongst others Niels Bohr – had the vision of a collective, European, nuclear research laboratory, spanning the borders, bringing the countries closer together and sharing the equipment costs. Thus the “Conseil Européen pour la Recherche Nucléaire” (Cern) was founded in 1954 by 11 different European governments. Nowadays it has grown to the largest center for particle physics research in the world.

Cern is financed by 20 European member states and has an annual budget of around 720 million Euro, but it also collaborates with many non member states all over the world. At the moment around 2500 people are employed by Cern to design, built and operate the experiments. Over 8000 scientists visit Cern regularly, among them half of the worlds particle physicists.

Currently the most powerful particle accelerator of the world, the large hadron collider (LHC) went into operation. It will clash lead ions or protons with a final energy of 7 TeV each, which allows to reconstruct the conditions very close to the big bang in a laboratory environment. The objective of the LHC is – amongst others – to prove the existence of the Higgs boson, one of the last standard model particles, that have not yet been observed in nature.

To realize the particle physic experiments, Cern has already pushed the technological boundaries of its age many times, resulting in spinoff technologies that entered everyday life. The most prominent example is the World Wide Web, initially developed in 1990 to have a distributed platform for the documentation of experiments. Parallels can be drawn, as the research of radio frequency to DC conversion – like discussed in this work – has a potential to be useful outside the world of particle accelerators too. Wireless powered sensor networks or even consumer products are imaginable.

1.2 Brief introduction to particle accelerators

1.2.1 How to accelerate particles with RF

How a charged particle is influenced by an electric or magnetic field is given by the Lorentz equation.

$$\vec{F} = q(\vec{E} + \vec{v} \times \vec{B}), \quad (1.1)$$

Only the electric field can be used for accelerating, as the force resulting from the magnetic field is always perpendicular to the velocity $\vec{v}$ and the magnetic field vector $\vec{B}$. Thus there is no energy gain possible by magnetic field alone. However, it is still useful in changing the particles trajectory.

Simple low energy accelerators can be built by using a static electric field $\vec{E}$. Oppositely charged electrodes exert a force on the particles and accelerate them. This principle has been used in the early 1920s by the accelerator pioneers John Cockcroft and Robert J. Van de Graaff which both became famous for their invention of very high voltage generators. The DC principle is still widely in use today: The Cathode Ray Tube in televisions or X-Rays tubes are essentially low energy electron accelerators utilizing an electrostatic field. The maximum achievable static voltage is the limiting factor on particle energy for those devices.

It was Rolf Wideröe – a PHD student of Karlsruhe university – who was the first to get beneath this limitation. In 1927 he constructed the first linear accelerator based on a time varying electrical field. A cylindrical electrode first attracts, and after its crossing, repels the particle bunch. This way the particles get a kick in velocity while crossing each electrode (from now on called drift tube) and can ultimately obtain an energy higher than the maximum voltage difference. The general principle can be compared to a linear synchronous motor. The phase of the driving voltage has to be synchronized to the position of the particle bunch, which means that with increasing velocity of the particles, either the frequency or the spacing of the drift tubes has to increase. The later, with increasing distance along the accelerators length, is usually done for linacs. The drift tubes of Cerns linac1 can be seen in Figure 1.1.

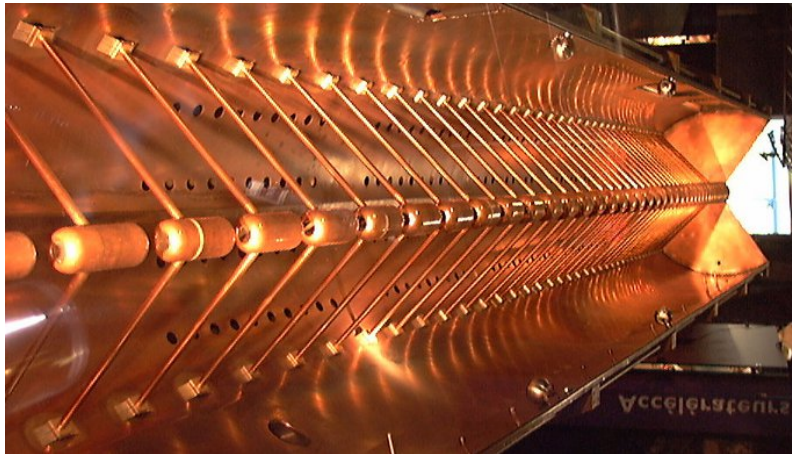


Figure 1.1: Drift tubes of linac1 at Cern.

Practical structures are only possible if the driving frequency reaches the Radio Frequency (RF) range. As the particles reach relativistic velocities, they do not gain much more speed. The velocity stays constant – very close to the speed of light – while the energy gain through acceleration increases the mass of the particles. As a consequence the accelerators can operate at a fixed frequency, the RF system can have a very small bandwidth. To minimize electromagnetic radiation losses, resonant RF cavities are built around the drift tubes. Depending if there is a reflected wave present in the cavity, two accelerating structures are distinguished: Travelling wave (TW) and Standing wave (SW) cavities. They differ substantially in their operation, although both need termination resistors that convert a big amount of the input RF power to heat.

1.2.2 Powering a cavity

The two cavity types distinguished in this work can be seen in Figure 1.2.

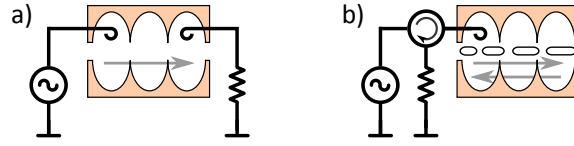


Figure 1.2: RF Architecture of a travelling (a) and standing wave cavity (b)

In a **travelling wave** structure, only the forward wave is present. The cavity behaves like a matched transmission line. The particles travel with the phase velocity on the crest of the forward wave. To ensure the travelling wave operation and inhibit any reflected power, the cavity has to be terminated by a matched resistor. The RF generator always sees the same load impedance and never any reflected power, regardless if there is a beam present or not. So additional amplifier protection features like circulators are not strictly necessary. Basically the power not ending up accelerating the beam is converted to heat in the termination resistor. As this fraction – depending on what kind of particles are accelerated – can reach 70 % and the cavities are energized nearly continuously, there is an enormous potential for energy recovery. The biggest TW based accelerator at Cern is the Super Proton Synchrotron (SPS) with an RF power of 2.6 MW [1].

A **standing wave** cavity operates by a different principle. RF power is coupled into the cavity at one side and reflected by the walls on the other side. The superposition of incident and reflected wave generates a standing wave pattern in the cavity. This standing wave creates the voltage gradient in the gaps between the drift tubes. To prevent deceleration after the gap, the drift tubes shield the particle from the wrong polarity part of the wave. In this architecture there are operating conditions in which the RF amplifier sees a short circuit. So generally a circulator is used to protect it against the reflected power. Reflection to the amplifier happens on the transient phase at RF power on, when the cavity starts to ring up (called the filling period) and during ringdown, when the RF power is cut. Between those transients, the cavity reaches a steady state, where no power is reflected. A potential for energy recovery can be seen especially during the transients. Accelerators with normal conducting sw. cavities suffer from large losses in the walls and often operate in pulsed mode. That means RF is only switched on when a particle bunch is about to enter the cavity. This way the cavity might be filled and left ringing down several thousand times a second. One solution to conserve energy, especially in this application is shown in Chapter 2.3. The LHC is based on SW cavities and operates with a maximum RF power of 3.5 MW [2].

1.3 Motivation for energy recovery

The motivation for this work is not to design and realize a working energy recovery system for the SPS, this would exceed the boundaries of this Thesis by far. It is rather meant as a feasibility study, giving a solid foundation for further work on the subject. The author expects the first semiconductor based recovery system deployed in 3 - 5 years.

After an overview on different methods of energy recovery is carried out in Chapter 2, this work focuses on the most feasible one, utilizing Schottky diodes. The problems arising with those devices in the RF band – particularly their parasitic capacitance – is shown and reviewed in Chapter 4. Resonant rectifiers can be built to compensate

this parasitic to some degree (Chapter 4.3). However, there are limits to this approach and although Schottky diodes with reverse voltages in the kilovolt range and forward currents in the tenths of amperes are now available¹, they would perform poorly in an RF rectifier at higher frequencies. The utilized semiconductor rather has to be carefully chosen by means of its power handling capability in relation to its capacitance and switching time (Chapter 5.4). To reach the design power, an array of those rectifier modules must be deployed, the number of modules depends on their individual power handling ability. The main driving force behind this research work is the fact, that semiconductor materials improved steadily over recent years, state of the art devices are nowadays powerful enough to make the rectifier array become economic and energy recovery feasible.

In this work, a conceptual design of an energy recovery system is presented. The design is tailored to the SPS, as this machine works on a relatively low frequency of 200 MHz and provides a big amount of energy that could be recovered. However the SPS is just used as an example, to show which engineering challenges will arise and how they could be solved.

1.3.1 Potential for the SPS

The SPS is a high energy proton accelerator and the pre injector for the LHC at Cern. It was designed as a highly flexible machine, accelerating lead ions, protons, antiprotons, electrons and positrons. Heavier particles need a bigger frequency swing to accelerate to the same energy², thus the SPS cavities must have a wide bandwidth. At reasonable power levels, regular SW cavities would necessity a variable tuning mechanism, while TW cavities can provide this bandwidth easily. For this reason and for the fact, that they always appear as a matched load and so can be connected by long transmission lines to the amplifiers, four TW cavities were deployed in the SPS [1]. Its main technical data is shown in Table 1.1.

The SPS [1]		One SPS Cavity	
Circumference	6.9 km	Min. operating frequency	199.526 MHz
Max. Particle Energy	450 GeV	Max. operating frequency	200.396 MHz
Average Beam Current	70 mA	Peak Voltage Gradient ³	1.8 MV
Number of Cavities	4	Energy gain for one turn	1.27 MeV
RF Architecture	TW.	Power dissipation in walls	13 kW

Table 1.1: Technical data of the SPS

Each Cavity is driven with up to 650 kW RF [3] at 200 MHz, produced by four Tetrode tube amplifiers. The amplifiers are located on the surface. A large diameter transmission line brings the RF power down to the tunnel, 60 m underground, where the accelerator is located. To permit the travelling wave operation, the cavity is terminated by two water cooled 50Ω power loads, each capable of dissipating 500 kW continuously.

¹Third generation Silicon Carbide Schottky diodes from infineon: <http://www.infineon.com/cms/en/product/channel.html?channel=ff80808112ab681d0112ab6a50b304a0>

²As lighter particles like electrons are already a lot closer to relativistic velocities with their injection energy. During acceleration they gain energy in the form of mass and hardly increase velocity.

³For one complete cavity with multiple accelerating structures.

The loads are located in the tunnel, directly above the cavity and can be seen in Figure 1.3.



Figure 1.3: SPS Cavity and Termination Load assembled in the tunnel.

The power to the termination loads P_L can be estimated by taking the known amplifier output power P_{AMP} and subtracting the losses from the transmission lines, cavity walls and the beam. For one cavity this is:

$$P_L = P_{AMP} - P_{TL} - P_{CAV} - P_{BEAM} \quad (1.2)$$

P_{AMP} The output power of the RF amplifier is controlled, depending on the active machine cycle. The most used one in the years 2009 and 2010 was “Cycle 2009”, which lasts 42 seconds, then it repeats. The peak power reaches 650 kW per cavity. The average power over one cycle is used for the calculations which is 254 kW.

P_{TL} The air filled transmission lines to the tunnel have a diameter of 35 cm and a length of 110 m. The transmission loss is given by $0.132dB/100m$ so 8.4 kW of power is dissipated.

P_{CAV} The cavity walls are not superconducting, so they will suffer from resistive losses. At full power, 13 kW is dissipated in the watercooled cavity walls [1].

P_{BEAM} The actual power going to the beam can be calculated from the particles energy gain per turn and the average beam current: $1.27MeV * 70mA = 88.9kW$. In the mentioned machine cycle, the beam is present for 47.4% of the time. In conclusion, in one cavity an average power of 42.1 kW is converted to an rise of particles energy.

Summing it all up, the average power to the termination load for one cavity and a typical machine cycle is $P_R = 190.5kW$ which is 75% of the input RF power. This power is converted to heat directly in the tunnel. Water cooling is used to move the heat out of the tunnel and dissipate it on the surface.

Just to show the potential and the motivation of this research work, it shall be assumed that 75% of P_R from all 4 cavities can be recovered and fed back to the utility grid (which is a feasible estimation, shown in Chapter 8) the results are:

- Cern saves an annual 5 million kWh in electricity.
- Cern saves over **335 000 €**⁴ in electricity costs each year.

Investing in the development of an energy recovery device would certainly pay off, help to protect the environment and also advance knowledge on RF/DC conversion, which is an expertise, many fellow operators of particle accelerators will be strongly interested in the future.

Another potential candidate of interest is the European Spallation Source, which is a linear accelerator currently built. Its construction is estimated to be finished in 2018. By incorporating energy recovery in the design, power consumption could be reduced and the plumbing in the klystron gallery simplified. The overall RF power is 3.8 MW. There are multiple RF systems at 352.2 MHz and 704.4 MHz [4]. Those frequencies are well within reach of an energy recovery system with semiconductors.

⁴Assuming 0.067€/kWh, which is the statistical electricity price for industry in France for November 2009. Source: <http://www.energy.eu/#industrial>

2. Possible methods for energy recovery

Technologies, suitable for making particle accelerators more efficient can be separated in two categories. Ultra high vacuum and semiconductor devices.

The first category is mainly found at institutions working with particle accelerators. Cavity resonators or electron beams can be used for energy conversion and storage.

The second category can also be found in more commercial areas of technology. Nowadays switched mode power supplies – operating at RF frequencies – and wireless powered devices like RFID pose a need for efficient RF/DC converters. Besides that a lot of fundamental research has been done on wireless power transfer, contributing to a design of a solar power plant, located in a lower orbit around the earth.

Part of this chapter was presented as a poster on the International Particle Accelerator Conference (IPAC) 2010. The results are summarized in the publication [5].

2.1 Direct feedback

One idea to recover energy in a travelling wave cavity is to get rid of the termination loads and directly feed the RF signal – with the right phase – back to its input. The phase of the fed back signal is adjusted so constructive interference happens and the cavity partly powers itself by its wasted power. This is done by using cables with an electrical length of $\lambda/2$. A microwave power combiner is used to sum up both signals.

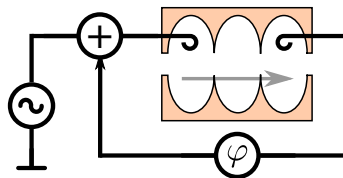


Figure 2.1: Unfeasible direct RF feedback.

The first problem is that the operation frequency and thus λ changes slightly while the particles gain energy. This change happens on a timescale of several seconds and might be easily compensated by a closed loop system that keeps the phase shift φ zero.

However, another fundamental problem arises from this approach. The proposal implicitly assumes, that the cavity behaves as a linear, time invariant system. The output signal must have the same frequency as the input and might be changed by a **constant** factor in phase or amplitude. Once a beam is present, this can not be assumed anymore. In fact the interaction of a relativistic particle bunch with the electromagnetic field creates certain transient responses at the cavities output. In fact, the cavities voltage is the convolution of the cavities impulse response with a rectangular function describing the bunch distribution over time [6]. During normal operation of the SPS, the spacing between two bunches is around 25ns, changing phase and amplitude response

on that timescale. This makes it impossible to feed the power back, always “with the right phase”, causing all kinds of interferences on the cavities input signal.

In this approach, the power combiner would have to be highly asymmetric, almost 70 % of the cavities input power would be delivered by the feedback loop. This means the actual RF source loses influence on the cavity voltage. Instead of strongly determining it, the RF source is now rather weakly coupled to a resonance structure. The voltage in the cavity is especially delicate, as three control loops – for amplitude, frequency and phase – are influencing the RF source to keep the beam stable. Adding another physical kind of feedback loop to this – causing all kind of distortions by transient phase errors – and coupling the actively controlled source weakly to it, adds a whole new dimension in system complexity and makes it simply impossible to keep the beam under control in real life.

That is the reason why a solution, converting the outgoing RF to another form of energy, not changing the behavior of the system, is preferred.

2.2 High temperature loads

The problem with classical termination loads is, that the temperature of the cooling water is only several Kelvin above ambient. This makes it very difficult to use the thermal energy technically. A solution would be to deploy special high temperature loads that operate with less flow rate but dissipate heat at $> 100\text{ }^{\circ}\text{C}$.

This way the thermal energy could be used to provide heating to the building or even to convert it back to electricity by a stirling engine or steam turbine. The latter is often used in solar power plants. Water or even molten salt at temperatures of $> 200^{\circ}\text{C}$ and pressures $> 50\text{ bar}$ is used as heat storage and exchange medium [7].

For small scale solar plants with power levels $< 5\text{ MW}$, a thermal to electrical efficiency of 19 % can be reached with a turbine setup and $200\text{ }^{\circ}\text{C}$ water temperature [8]. At higher temperatures the turbine will operate more efficient [7].

Advantages of this approach are the minimum changes that have to be done to the actual RF system of the cavity in the tunnel. There are no problems with RF reflections or harmonics as the load resistor stays in place. The disadvantage is the quite low efficiency but high complexity and space requirements of the steam turbine and generator system. Also high temperature RF load resistors are not available on the market yet.

2.3 Energy storing cavities

In most non superconducting cavities, the energy losses are too big for CW operation. One solution for this, is to switch the RF on, just before the particle bunch arrives. Unfortunately standing wave cavities have a certain transient response and it is this filling time, where most of the power is lost in termination loads. So especially if the time between particle bunches is short in relation to the filling time, CW operation is the only option. A new particle bunch would arrive even before the transient response could decay.

For exactly this reason, the predecessor of the LHC, the Large Electron-Positron Collider (LEP) was equipped with special energy storing cavities. Those were normal,

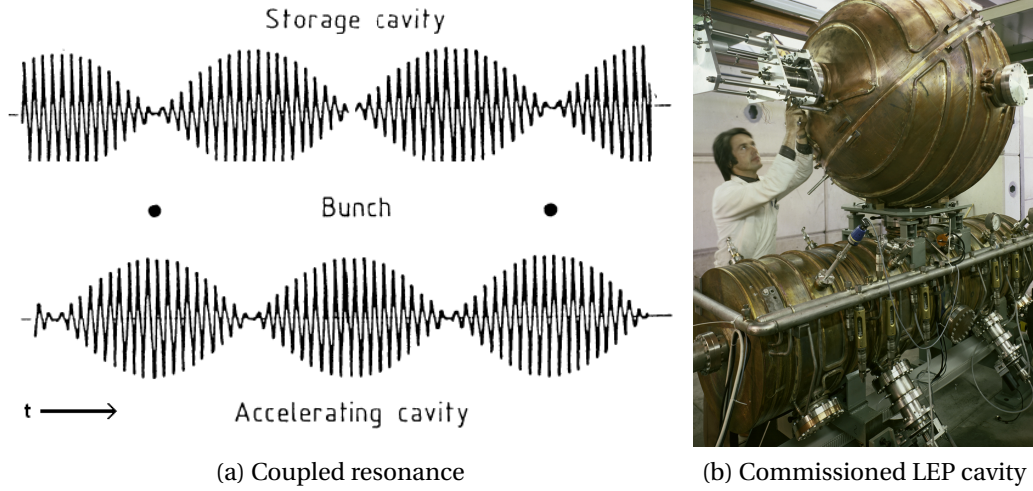


Figure 2.2: The LEP energy storage cavity

standing wave types, but utilized a special, high Q, low loss storage cavity, which cached the electromagnetic energy in between bunches [9]. The concept can be seen in Figure 2.2.

The main and storage Cavity have the same resonant frequency and form a system of 2 coupled resonators. Energy swings back and forth between the two cavities. Just as a bunch of particles enters the accelerating cavity, the electromagnetic field is at a maximum. The beat frequency of the system is adjusted to the bunch repetition rate. The advantage of this is, that the dissipation in the accelerating cavity, where the losses are high, is halved. The accelerating cavity is periodically filled and emptied but as the amplifier sees the whole system, it operates in CW mode. There are no transient responses coming from the cavity like in a traditional pulsed system and so considerable less reflected power needs to be dumped in resistors. Unfortunately accelerators equipped with energy storage cavities loose their flexibility. Also the system can not be easily retrofitted to existing systems. Energy recovery with an intermediate DC voltage is preferred as it does not change the systems operational behaviour, so doesn't cut back on the accelerators flexibility and can be easily retrofitted.

2.4 Cyclotron Wave Converter

The Cyclotron Wave Converter (CWC) is a new kind of RF to DC converter, patented by V.A. Vanke in 2003 [10]. Its basic idea is to use RF to accelerate a high current electron beam, then capture the electrons which will generate a DC voltage that can be used technically.

The schematic of a CWC device can be seen in Figure 2.3. An electron gun on the left provides a continuous, high current beam of electrons. They enter a classical cyclotron structure which is excited by the RF input power. The electrons move perpendicular to

¹The relation, of storing electromagnetic waves in a cavity, to the german tale of the “Schildbürger” – trying to catch light in buckets to illuminate their windowless city hall – shall be pointed out.

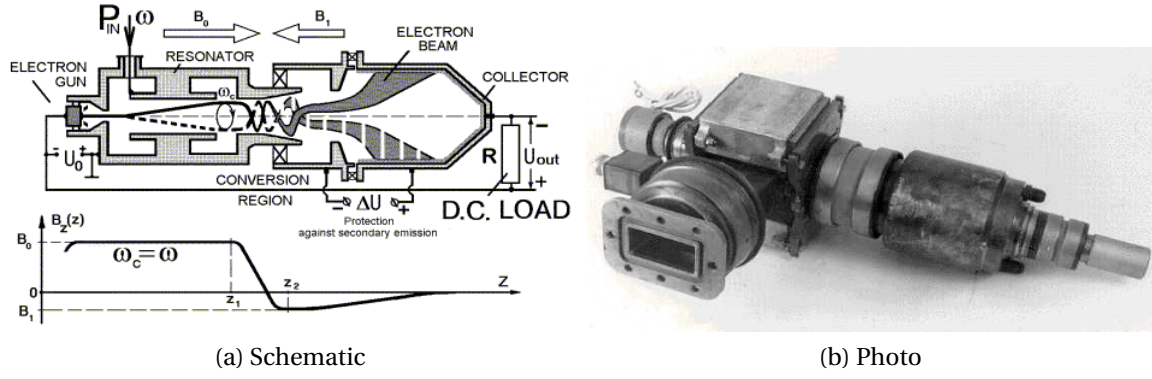


Figure 2.3: The Cyclotron Wave Converter

the uniform magnetic field B_0 and accelerate on a spiral path. A cyclotron is essentially a simple kind of particle accelerator, the RF energy is converted to rotational energy of the electrons.

The conversion region, in the middle of the device, shows a reversal of the magnetic field. The rotational energy of the electrons is converted into an increase in longitudinal velocity. A “depressed collector” catches the electrons and converts their kinetic energy to a large DC voltage at the load resistance.

The feasibility as an energy recovery device for the SPS is summarized in Tab. 2.1. A positive aspect is the large input power per unit as well as the ability of the device to stand large peak powers. The wasted power of the SPS could be recovered with only a few devices, keeping the system complexity low and preventing losses through excessive deployment of RF power splitters. The device shows high power ratings while still converting RF to DC with good efficiencies, making it suitable for the application.

However, the main disadvantage of the CWC – for this application – is its need for a resonant cyclotron RF-Cavity. At the operating frequency of the SPS, which is 200 MHz, this would yield an exceptionally large device. The creation of the required magnetic field gradient on this scale would be quite an engineering challenge. The other disadvantages are the need for an ultra high vacuum, the maintenance requirements and the warm-up time.

Property	Value	Suitability
Power range [kW]	0.5 ... 50	✓
Achieved efficiency [%]	83	✓
Output voltage [kV]	10 ... 100	⚠
Operating frequency [GHz]	1 ... 50	⚠
Usable Bandwidth [%]	0.5 ... 5	✓

Table 2.1: Positive and negative aspects of the CWC

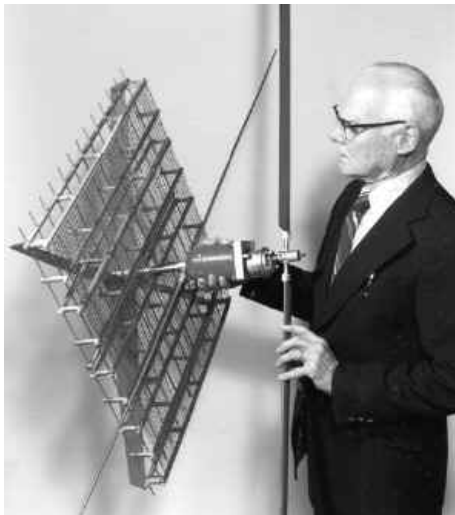
In [10] the CWC device has been implemented and tested for the recovery of RF Energy in a particle accelerator. In a simulation the authors were able to predict an overall RF to DC efficiency of 73% in the S-Band at an input power of 1 MW. This could not be confirmed in measurements because the electron collector broke down as the output

voltage reached 125 kV. At these voltages, considerable engineering effort is needed to prevent losses through corona discharge, breakdown or even danger by the generation of X-Rays.

This shows that the CWC device is not easily scaled to higher power levels, as the high electron energy and the consequent high output voltage poses a limitation.

2.5 Rectennas

The idea of using semiconductors to convert RF-Power to DC has been around since 1964 when W. C. Brown invented the rectenna. He was able – with this combination of antenna and diode-rectifier – to transmit a substantial amount of power wirelessly. One of the first prototypes operated at 2.5 GHz and used an array of 4480 small signal point contact diodes. The DC output power was 270 W, enough to build a demonstration helicopter – shown in Figure 2.4a – which held itself in the air by RF-Power only [11].



(a) Mr. Brown & his helicopter.



(b) The Goldstone experiment.

Figure 2.4: Microwave power transmission

At this point, Nasa got involved in the project, aiming for a solution to transmit large amounts of solar power from space to earth. For this undertaking, the name Solar Power Satellite (SPS²) is well known in Literature.

With the financial support from Nasa, a second demonstration of this technology was done in 1975 at the Goldstone Complex in the Mojave Dessert. The setup can be seen in Figure 2.4b. A large scale rectenna produced over 30 kW of DC power with an RF/DC efficiency of 84% at 2.45 GHz. The microwave power was transmitted over a distance of 1 mile. The rectenna was constructed as thin film etched circuit. Each element consisted of an input filter to keep internally produced harmonics from being radiated, a small signal Schottky diode and a DC output filter. The elements are connected in series strings which in turn are connected in parallel to power the load [11].

²Not to be confused with the Super Proton Synchrotron (SPS) at Cern.

The main drawback of rectenna like devices is the very low power handling capability of the single elements. The most powerful element found in literature could handle 6 W [12], this is adequate for microwave power transmission – big rectenna arrays receive microwaves with low power density and their individual DC outputs are combined. However, this is not suitable for an SPS termination load, where RF power levels of several hundreds of kW need to be rectified in a small space. The rectenna concept needs to be improved and at least 200 W must be processed by a single device, else the recovery system would not be economical. Especially the power splitter would be too lossy and would occupy too much space.

Working with modern semiconductors and a much lower operating frequency will allow to build an RF/DC converter – based on an improved rectenna – that reaches these power levels.

2.6 Radio frequency DC/DC converters

Research on DC/DC converters is especially interesting, as each device also contains a rectifier, contributing strongly to the overall system efficiency. Contemporary research in DC/DC converters is mainly motivated in increasing the devices power density. This is achieved by extensive integration and miniaturization of components. The only way to reach this for passive components is to increase the operating frequency. With a higher frequency, less energy storage is needed and smaller inductors and capacitors can be used. Also the step response of the converter is much faster, which is important for fast output voltage regulation. Nowadays commercial DC/DC converters already operate in the MHz range and research has been done on high frequency converters operating in the hundreds of MHz³. Fast rectifiers are also needed for RFID - like technologies, that are powered by electromagnetic radiation.

Year	Reference	Frequency	Power	Efficiency
1980	[13]	10 MHz	5 W	> 68 %
1988	[14]	22 MHz	50 W	88 %
1988	[15]	22 MHz	50 W	> 78 %
1993	[16]	100 MHz	40 W	38 %
2002	[17]	100 MHz	3 W	80 %
2006	[18]	30 MHz	200 W	> 87 %
2006	[19]	100 MHz	10 W	> 75 %

Table 2.2: Literature on RF/DC converters.

An overview of the published results from interesting rectifiers built in laboratories can be seen in Table 2.2. Some papers only mention the overall efficiency of the whole DC/DC converter, the rectifier itself is surely more efficient. Those are indicated by a > symbol.

Taking advantage of the know how from rectennas and RF switch mode power supplies, it is possible to build an RF/DC converter, which can handle power levels in the 100 W

³The potential misleading terms “VHF” or “UHF” are sometimes used in literature about power converters. Only some of those devices do operate in the *very high frequency* (30 MHz - 300 MHz) – and usually none in the *ultra high frequency* (300 MHz - 3 GHz) band. The authors of those publication merely want to emphasize the increased operating frequency compared to ordinary power converters.

range, avoids excessive losses through power splitters and the drawbacks of ultra high vacuum devices. This provides the foundation for the energy recovery concept, which is presented in the following sections.

3. Overview of the system concept

The energy recovery system is intended to replace a matched high power, RF termination resistor, meaning it should behave as such.

- Its input impedance must be purely resistive, constant, independent of power and have a value of 50Ω . No significant amount of power – at any frequency – is allowed to be radiated on its input.
- It must be reliable and fail safe under all operating conditions. If there is no demand, the recovered energy has to be dumped in resistors to allow classical, lossy operation.
- A defect in one component must not lead to a failure of the whole system, instead a graceful degradation behavior must be achieved. Damaged parts must be easily exchangeable.
- The recovered energy must be available in a form, easily fed back to the 50 Hz utility grid or used on the DC rails of the RF amplifiers.
- Its recovery efficiency must be as high as today's technical limits allow it to be.
- The modules should be as simple and cheap to manufacture as possible. This is especially important as several hundred modules will be needed and the probability of a failing module is directly related to its complexity and component count.
- Preferably it should fit in the available space in the tunnel, also the cooling requirements must then be fulfilled there.

To meet all these requirements, the recovery system is proposed as shown in Figure 3.1. In this chapter, each element will be introduced, its purpose and technical difficulties explained.

3.1 RF - Power splitter

It has been shown in Chapter 1.3.1, that an typical average power of 191 kW goes to the two termination loads. However, the recovery system needs to be designed to handle the peak power of up to 650 kW. As no single semiconductor device can handle those power levels, both of the termination loads are replaced by coaxial power dividers. They spread their input power of 325 kW into several hundreds of channels of 1 kW each. One divider consists of a modified 50Ω transmission line with inserted coupling pins. Those act like an antenna, providing a small part of the input power, depending on its length and geometry. As the field density in the transmission line will diminish along its length, the coupling of subsequent antennas has to increase for

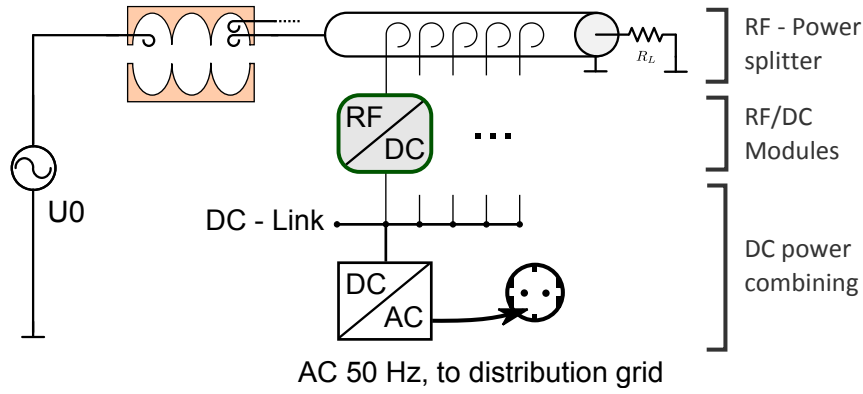


Figure 3.1: Architecture of the RF - Recovery device.

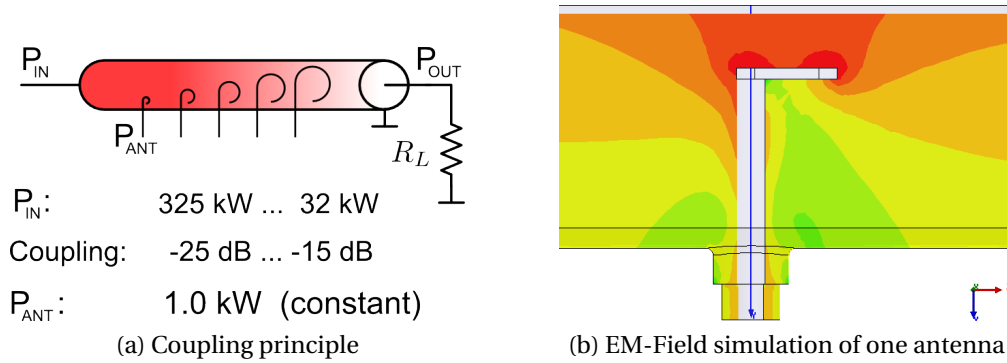


Figure 3.2: The coaxial power divider.

constant output, shown in Figure 3.2a. For economical reasons, the transmission lines are terminated by a resistor R_L at its end where 10 % of the input power is absorbed.

The coupling values have been obtained and the feasibility checked by an electromagnetic field simulation [5], the results are shown in Figure 3.2b. An air filled transmission line with an inner conductor diameter of 100 mm has been chosen. A capacitive coupling geometry turned out to be the most feasible¹.

In this case 325 kW is coupled out to the rectifier circuits in 33 stages. Each stage, except the last, has 10 coupling antennas distributed around the circumference of the outer conductor of the coaxial line. As the antennas need to have a bigger size towards the end, the last stage consists only of 5 antennas. The overall length is about 3.3 m. Further investigations are ongoing to reduce the length of the system.

Each antenna is optimized to couple out 1kw of RF power into a 50 Ohms impedance. They are connected by a short coaxial cable to the RF/DC Modules.

3.2 RF/DC modules

The RF/DC modules are the heart of the recovery system. Besides performing the conversion in the most efficient way, they contain additional components to ensure a fail

¹Credit for designing the power divider and accomplishing the EM simulations goes to Alexj Grudiev.

safe and low reflection operation. They are designed for an input power of 1 kW, coming from the power divider. This number has been chosen as it gives a good compromise between losses in the coaxial power divider, system complexity, space requirements and price of the RF components.

Worst case, the modules will dissipate up to 1 kW of power. Deployed in a large array, a vast amount of heat will be produced in a small space. To simplify cooling, the RF/DC modules should be built with all power components (diodes, RF loads) on a common copper plate. Deployed in a large array, the modules will be mounted on metal structures with integrated cooling water pipes. Heat can be efficiently removed while the internal construction of the module stays simple. This method is well-proven and has been used in the solid state amplifier of the Soleil synchrotron [20].

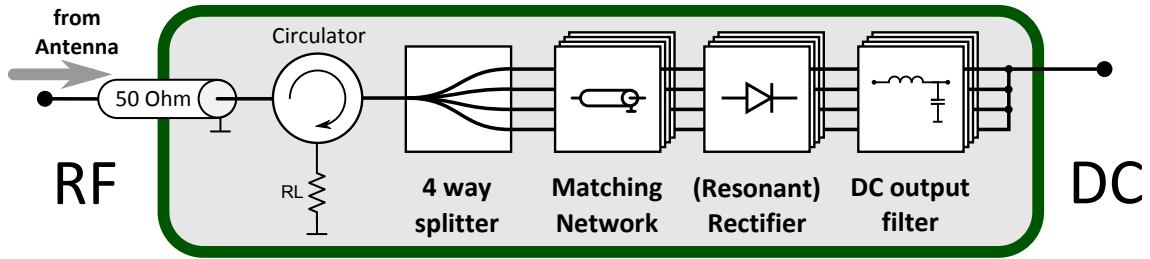


Figure 3.3: The proposed RF/DC Module

Circulator The rectifier itself is a nonlinear device and will show a differing amount of reflections for different input powers. As power is determined by the active machine cycle and changing over time, reflected power can not totally be prevented. Also if the rectifier fails, it is likely to show a short circuit on its input [21] and reflect all of the incident power.

For those cases, the circulator directs the reflected power into the backup resistor R_L . It needs to be rated for the full nominal module power of 1 kW and might be cooled by the already existing water plumbing in the tunnel. This way, the rectifier modules become fail safe, under every possible operating condition. Even for an rectifier failure, no RF is reflected on the module's input. The overall energy recovery system will show a graceful degradation, the recovery efficiency might degrade with each failure, but the cavity is still terminated correctly.

For cost reasons, only a narrow band circulator is economical. Power on harmonic frequencies – radiated from the rectifier – might be able to pass the circulator and would be transmitted towards the Cavity. This needs to be prevented and is the reason why much effort is done, designing a resonant rectifier, which does not radiate substantial power on harmonic frequencies.

4 way splitter It showed in Chapter 7, that a single rectifier can not handle the full 1 kW RF power, supplied by the coaxial power divider. One rectifier is rated for a nominal power of 250 W, so four of them are needed.

One solution would be three microwave hybrids, cascaded in two stages. The 1 kW input power will be split to 2 x 500 W with the first hybrid and its outputs will be split to 4 x 250 W with the second and third one. To tolerate one rectifier failure, every hybrid needs a load resistor on the isolated port, rated for at least quarter of its input power. In the event of a failed rectifier, a quarter of the hybrids

input power will be reflected, a quarter will be dissipated in the isolation load and the remaining half will be converted to DC by the other rectifier. Heat dissipation is not a problem, as the modules will be water cooled, thus the isolation loads can be made relatively small in size. The circulator on the module's input is still needed to terminate the remaining reflected power.

More powerful semiconductors are expected to be available in the near future. As a single rectifier with 1 kW nominal power will become feasible, the 4 way splitter can be omitted, considerably reducing module complexity.

Matching Network It provides an optimum power transfer, while not exceeding the maximum voltage or current ratings of the diode. The small bandwidth of the input signal allows the use of standard resonant matching techniques utilizing LC circuits or transmission lines. Care must be taken with the design and simulation procedure, as the rectifier is a nonlinear device and its input impedance (in the range of 15Ω) is likely to change with different input power levels. A solution for impedance matching is found in Chapter 4.4.

Rectifier It converts RF to DC in the most efficient way, at the highest power possible with available semiconductors. Different rectifier architectures exist, dependent on number of diodes, diode stress, input / output impedance and harmonic radiation. They are compared in simulations (Chapter 4), the most suitable ones are chosen for a prototype. By exploiting resonance with the diodes junction capacitance, many advantages are obtained Chapter 4.3. Also the most suitable diode has to be found, device parameters like material, band gap and geometry of the diode chip have to be considered. Also active rectification with mosfets or Schottky drain transistors might be considered Chapter 4.6.

DC - Output The output filter provides a clean, RF free, DC output. Depending on the filter, the rectifier behaves as a current or voltage source. At RF frequencies, parasitic components of the passives present a major design challenge (Chapter 7). It should be noted, that the optimum efficiency is only reached if the DC load resistance has a specific value, depending on the RF input power level.

3.3 DC power combining

The DC output from a large number of sources has to be collected and combined to an intermediate DC power rail. In doing so, the rectifiers should see a certain optimum load resistance – which is variable and depends on the RF input power – otherwise the RF/DC efficiency will suffer. Commercial solutions already exist to solve this exact kind of problem, namely power converters for large photovoltaic arrays.

As the RF/DC modules essentially behave like a current source, they are especially suited for series parallel connection. Modules are connected in series strings to reach a usable voltage level (300 V - 1000 V), those strings are connected in parallel to sum up the current and provide redundancy in the case of device failure.

It is more likely for the modules to fail with the output shorted [21]. In this case the current in the series string can still flow and the overall system operation is – despite the slightly decreased recovery efficiency – not affected.

As the design of the RF power divider makes sure, that every RF/DC module provides the same amount of output power, very little power is lost in the combination through mismatch of a single module. This is a major problem in rectenna and solar cell arrays with a large number of series devices. The overall power drops considerably, if only one element performs weak or is not illuminated [22].

The combined DC power is fed to a commercial photovoltaic power converter. Thus two technical challenges are solved with minimum effort:

- The power can be fed back to the 50 Hz utility grid. Those devices are approved by power companies, contain the necessary safety interlocks and are designed for reliability and energy efficiency.
- The RF/DC module shows the best recovery efficiency with a specific load resistance, which in turn depends on the input RF power. Solar power converters are able to change this load resistance dynamically and incorporate algorithms to always operate at the maximum power point (MPP). This way, the rectifiers always reflect a minimum of RF, independent of power level and a maximum recovery efficiency is reached.

Alternatively, the recovered power could be fed directly to the DC - link voltage of the Tetrode power amplifiers that provide the RF for the SPS cavities. They operate at 8.5 kV and 10 kV DC [23], so a custom power converter, to step up the voltage and to provide safety features is necessary. This way the recovered power would be routed to the consumer more directly, avoiding unnecessary transformation stages in between, straightforwardly increasing the overall operating efficiency of the SPS.

4. Rectifiers

Most of the requirements, defined in Chapter 3 can directly be applied to the actual rectifier circuit. Its power rating determines the number of rectifiers that have to be used for the given input power, thus the complexity and size of the overall array and in the end, if the energy recovery system can be realized in an economic way.

In Chapter 4.1, a simple rectifier circuit is analyzed analytically. Different diode parasitics are shown and introduced in the circuit, one at a time in Chapter 4.2. This changes the circuits operation considerably at RF frequencies, the waveforms barely resemble those of the simple ideal rectifier.

Subsequently solutions are found to compensate some of those unwanted elements, resulting in resonant operation Chapter 4.3. Its advantages – among them the sinusoidal input current – are shown. Then ways are found, to transform the rectifiers input impedance to 50Ω (Chapter 4.4).

To answer the question, if a more suitable rectifier architecture than the analyzed one exists, different circuits from literature are compared, figures of merit are derived from PSpice simulations and the most suitable rectifier is chosen (Chapter 4.5).

4.1 General analysis of a rectifier

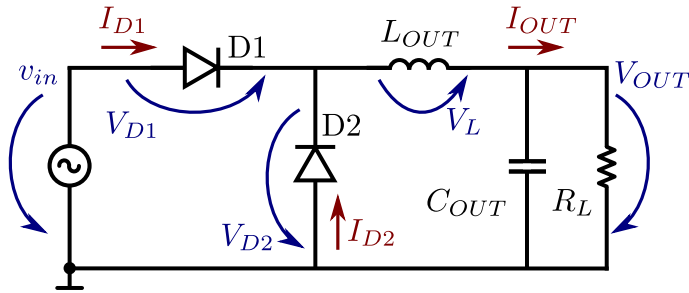


Figure 4.1: Schematic of a simple current output half wave rectifier.

A simple half wave rectifier [24], with a constant current output is shown in Figure 4.1. It is driven by an sinusoidal alternating voltage source v_{in} with an amplitude of $\hat{V}_{in}$ and a frequency of ω .

$$v_{in} = \hat{V}_{in} \cdot \sin(\omega t) \quad (4.1)$$

the voltage and current waveforms are shown in Figure 4.2.

L_{OUT} and C_{OUT} form a second order low pass filter. Its time constant is assumed to be so large, that the output current I_{OUT} appears constant (shown with a small ripple¹ in Figure 4.2). That means L_{OUT} forces a continuous output current through the load resistor, which flows through D1 on the positive cycle and through D2 on the negative

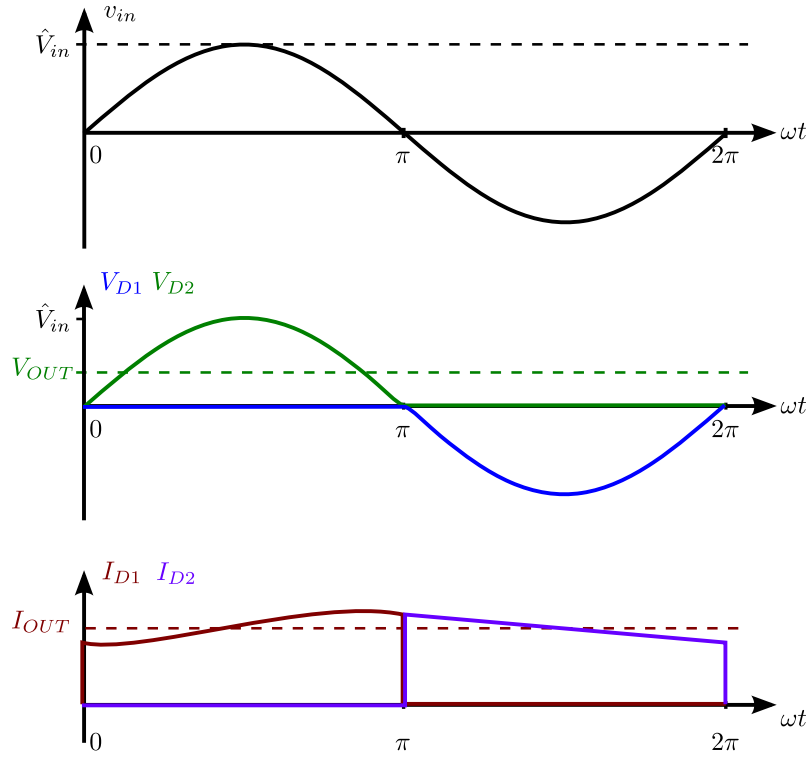


Figure 4.2: Waveforms of a current output half wave rectifier.

cycle of the input voltage. As input current is only flowing – and power is only delivered to the storage inductor L_{OUT} – when D1 conducts, this type of rectifier is referred to as half wave rectifier.

The diodes are assumed to be ideal switches, that means they do not have any voltage drop, reverse current or junction capacitance. To get a formula, relating in and output voltage, we follow the voltage loop of the output filter.

$$-V_{D2} + V_L + V_{out} = 0 \quad V_{D2} = \begin{cases} \hat{V}_{in} \cdot \sin(\omega t) & 0 \leq \omega t \leq \pi \\ 0 & \pi < \omega t \leq 2\pi \end{cases} \quad (4.2)$$

As the circuit is assumed to be in steady state, the average voltage across the inductor L_{OUT} needs to be zero,

$$\frac{1}{2\pi} \int_0^{2\pi} V_L(\omega t) d\omega t = 0 \quad (4.3)$$

Inserting Equation 4.2 into Equation 4.3 and solving gives an expression for the output voltage:

$$V_{out} = \frac{\hat{V}_{in}}{\pi} \quad (4.4)$$

This means the DC output voltage is directly proportional to the input RF amplitude.

¹The waveform of the current ripple can be calculated by considering the voltage across the inductor. In the first half cycle it is $V_L = v_{in} - V_{out}$, having a sinusoidal shape. The current through the inductor calculates as $I_{out} = \int V_L / L d\omega t$ resulting in an inverted cosine like waveform. During the second half cycle, there is a constant voltage across the inductor, resulting in a linear current decrease. As the inductor is assumed very large, in the following sections the waveform is assumed to be a flat square wave.

The input current to the rectifier is the same as I_{D1} . It is a rectangular waveform with a DC offset. Its amplitude is $I_{out}/2$. A rectangular waveform contains a lot of harmonics, which will be radiated towards the source. This poses a particular problem to the rectifier which will be treated in Chapter 4.3.5. Using the Fourier series, I_{D1} can be split into its frequency components. Each one can easily be denoted by using the harmonic index, which equals zero for the dc contribution, one for the fundamental, and two for the second harmonic.

$$I_{D1(0)} = \frac{I_{out}}{2} \quad (4.5)$$

$$I_{D1(1)} = \frac{I_{out}}{2} \frac{4}{\pi} \sin(\omega t) \quad (4.6)$$

$$I_{D1(2)} = \frac{I_{out}}{2} \frac{4}{3\pi} \sin(3\omega t) \quad (4.7)$$

$$I_{D1(3)} = \frac{I_{out}}{2} \frac{4}{5\pi} \sin(5\omega t) \quad (4.8)$$

As the input voltage is sinusoidal and has no overtones nor DC offset, effective power

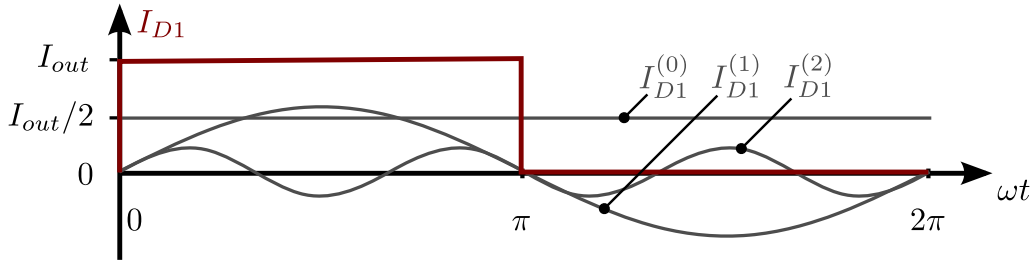


Figure 4.3: Frequency components of the input current I_{D1}

is only transferred by $I_{D1(1)}$ (for every other frequency, including DC, the average of voltage times current over one period is zero). The loss through overtones is described by the Power Factor as a ratio of real power by apparent power of the harmonics:

$$P.F. = \frac{P_{\text{real}}}{P_{\text{apparent}}} = \frac{RMS(v_{in}) \cdot RMS(I_{D1(1)})}{RMS(v_{in}) \cdot RMS(I_{D1})} \quad (4.9)$$

As v_{IN} and $I_{D1(1)}$ are shaped sinusoidal, their RMS value is obtained by division with $\sqrt{2}$, the RMS value of I_{D1} is

$$RMS(I_{D1}) = \sqrt{\frac{1}{2\pi} \int_0^\pi I_{out}^2 d\omega t} = \frac{I_{out}}{2} \quad (4.10)$$

So this gives a power factor of

$$P.F. = \frac{(2I_{out})/(\pi\sqrt{2})}{I_{out}/2} = 0.9 \quad (4.11)$$

This means, the maximum efficiency can not exceed 90%, limited by the non sinusoidal input current. 10 % of the input power is lost through harmonic radiation. In Chapter 4.3 ways are shown to avoid this limitation.

Another important measure – which will be used as a figure of merit to compare rectifier architectures in Chapter 4.5.2 – is the total harmonic distortion (T.H.D.). It gives

a hint, how closely the input current resembles a sine wave. It is defined as factor of power on harmonic frequencies to the power of the fundamental and related to the (distortion) power factor by

$$T.H.D. = \sqrt{\frac{I_{D1(2)}^2 + I_{D1(3)}^2 + \dots}{I_{D1(1)}^2}} = \sqrt{\frac{1}{P.F.^2} - 1} = 0.483 \quad (4.12)$$

The input resistance is important for matching the rectifier to the source, which – in the real system – will be the power divider with a generator resistance of 50Ω . As the source only transfers power on the fundamental frequency, the input resistance at this particular frequency is of interest:

$$R_{in(1)} = \frac{\hat{V}_{in}}{I_{D1(1)}} = \frac{\hat{V}_{in} \cdot \pi}{I_{out} \cdot 2} = \frac{V_{out} \cdot \pi^2}{I_{out} \cdot 2} = R_L \cdot \frac{\pi^2}{2} \quad (4.13)$$

It shows that the input resistance of the rectifier is directly proportional to the load resistance R_L .

The power capability factor (CPF) is an important measure, estimating how much power the rectifier can provide for given diodes. It is the maximum output power of the rectifier divided by the peak stress on the diodes. The maximum reverse voltage is

$$\hat{V}_D = \hat{V}_{in} = V_{out} \cdot \pi \quad (4.14)$$

the maximum forward current is

$$\hat{I}_D = I_{out} \quad (4.15)$$

So the power capability factor can be determined by

$$PCF = \frac{P_{rect}}{P_{diode}} = \frac{V_{out} \cdot I_{out}}{\hat{I}_D \cdot \hat{V}_D} = \frac{1}{\pi} = 0.32 \quad (4.16)$$

It will show in Chapter 4.3 that resonant operation will make the output power capability even worse.

4.2 Real life rectifiers

Unfortunately there is no such thing as an ideal diode and the parasitic elements of real diodes change the rectifiers operation considerably at RF frequencies. This section gives an overview of the most important parasitics and their impact on the rectifier.

Summing it all up, 4 different modes of operation – depending on which of the parasitic elements are considered – can be distinguished:

$L_p = 0, c_J = 0$ The ideal case has been analyzed in the beginning of Chapter 4.1 and results in pure square waves for the diode current.

$L_p = 0, c_J = \text{big}$ The analysis with consideration of the junction capacitance can be found in Chapter 4.2.2. The capacitors are charged and discharged 2 times each cycle. The diode current is overlaid by the differentiated input voltage.

$L_p = \text{small}$, $c_J = \text{small}$ What happens when a small package inductance is introduced is explained in Chapter 4.2.3. The circuit is operating the same as in the previous case, only now the square diode current is overlaid by high frequency ringing from the LC circuit. Characteristic for this case is that the time constants of the LC circuits are much smaller than half of a period of the input RF voltage.

$L_p = \text{big}$, $c_J = \text{big}$ As the time constant of the LC circuits approaches half a period time, we can not talk of ringing anymore. Rather the resonant frequency of the LC circuit is synchronized to the rectifiers input voltage. This results in many advantages: Current flow becomes sinusoidal, high frequency currents are limited and energy from the parasitics is recovered by the class E resembling resonant operation. It is explained in Chapter 4.3.

The main parasitics of a diode are shown in Figure 4.4.

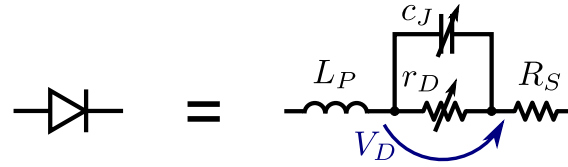


Figure 4.4: Simple equivalent circuit of a real Schottky diode.

The static parasitics R_S and L_P are caused by the bonding wire, the package pins and the circuit traces going to the diode. In addition the resistance of the semiconductors crystal lattice is not zero, especially for Schottky diodes, increasing R_S . The junction capacitance c_J in combination with the package inductance L_P is especially troublesome. L_P does not allow a quick change of current, which is needed to recharge c_J as the diode goes from forward to reverse operation. L_P can be up to 30 nH for standard packages like TO-254 or TO-3 and around 3 nH [25] for special RF packages like those shown in Figure 5.10a.

4.2.1 Static characteristic r_D and R_S

Any diode is a component with an inherently nonlinear relationship between voltage and current. In the equivalent circuit, the V/I characteristic of the diode junction is represented by r_D . This resistance depends on the voltage V_D across the junction, models the actual rectifying character of the diode and can be described by Equation 4.17, discovered by William Bradford Shockley. This equation is valid for forward and reverse operation, as long as avalanche breakdown effects of the semiconductor are not considered.

$$I_D(V_D) = I_S \left(e^{\frac{V_D}{nV_T}} - 1 \right) \quad r_D = \frac{V_D}{I_D(V_D)} \quad (4.17)$$

In most datasheets, the forward voltage is drawn in a diagram over the forward current (like in Figure 4.5). Then r_D can be seen as the inverse of the tangent in one point on the curve and can be determined graphically.

Parameter	Description
I_S	Reverse bias saturation current. Defines the current that flows because of minority carrier recombination, when the diode is in reverse condition. It is nearly constant over voltage and heavily dependent on temperature.
n	Ideality factor, varies between 1 and 2, depending on the kind of conduction process in the semiconductor device. For $n=1$, the device conducts current only by minority carrier diffusion (like in a Schottky diode). For $n=2$, the device conducts current only by recombination of holes and electrons in the junction. Values in between are possible (like for PN diodes).
V_T	Thermal voltage. $V_T = k \cdot T / e$ where k is the Boltzmann constant, T the absolute temperature in Kelvin and e the elementary charge. $V_T = 26mV$ at room temperature (300 K).

Table 4.1: Parameters of the diode equation.

The parameters for a GS150TC25110 Schottky diode can be found in Table 4.2. Its static forward characteristic can be seen in Figure 4.5. For the blue trace, only r_D has been considered, ignoring the static series resistance. The red trace also considers R_S and shows that this parasitic needs to be taken into account for Schottky diodes, it changes the devices characteristic considerably.

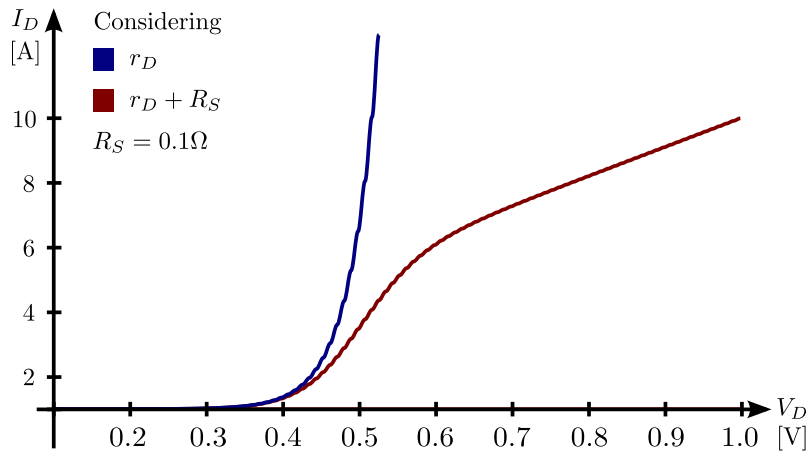


Figure 4.5: Static characteristic for a GS250TI25110 Schottky diode

Power is dissipated in the diode by r_D and R_S , the amount can be determined by measuring the area under the curves in Figure 4.5. Schottky diodes have – compared to PN devices – a lower forward voltage but higher series resistance (see Chapter 5.2). Thus at higher current levels, the losses in R_S dominate. If the diode is used in a resonant rectifier at RF frequencies, there will be a vast amount of purely reactive currents flowing through the junction capacitance. They will cause power dissipation in R_S but not in r_D .

For a Schottky diode the forward and reverse losses are determined by the barrier height. Higher barriers mean bigger forward resistance R_S but also less reverse leakage I_S [26].

It is the nonlinear characteristic of the diode that makes it rectifying but also radiating on a whole bunch of harmonic frequencies. This can be shown by performing a Taylor approximation of the diodes characteristic, resulting in a third order equation:

$$I_f = \alpha \cdot V_D + \beta \cdot V_D^2 + \gamma \cdot V_D^3 \quad (4.18)$$

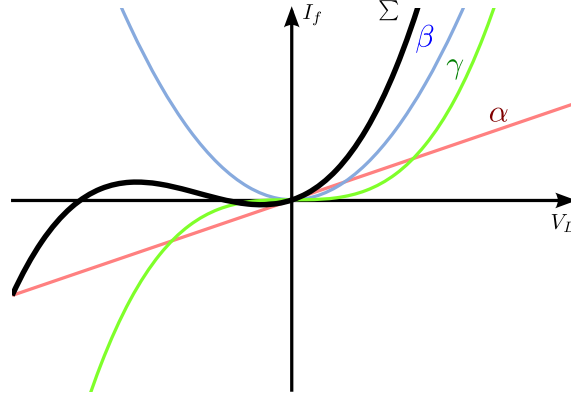


Figure 4.6: The 3 terms of the Tailor approximation of the diodes characteristic.

The diode is driven by a sinusoidal voltage source

$$V_D = \hat{V}_{in} \cdot \sin(\omega t) \quad (4.19)$$

V_D is inserted in Equation 4.18. It can be simplified by addition theorems and the resulting parts of the equation can be sorted by its frequency components:

$$\begin{aligned} DC: & \quad \beta \cdot \hat{V}_{in}^2 \cdot 0.5 \\ \omega: & \quad \alpha \cdot \hat{V}_{in} + \gamma \cdot \hat{V}_{in}^3 \cdot 0.75 \\ 2\omega: & \quad -\beta \cdot \hat{V}_{in}^2 \cdot 0.5 \\ 3\omega: & \quad -\gamma \cdot \hat{V}_{in}^3 \cdot 0.25 \end{aligned} \quad (4.20)$$

It can be seen that the waveform of I_f now contains new harmonics, not emitted by the voltage source but created by the diode itself. This behavior can only be observed in nonlinear systems. In fact the DC part and thereby the rectification property of the diode depends only on β , thus on how much the diode characteristic resembles a parabola.

The frequency parts 2ω and 3ω are the harmonics. When a sinusoidal voltage is applied to the diode, it no longer stays sinusoidal but is changed to a half wave rectified sine waveform. Looking at the signal in the frequency domain, we see those additional spectral lines appearing at harmonic frequencies. The diode will turn to a source, radiating power on these frequencies. It depends now on the surrounding circuit, the diode is embedded in, if those harmonics are suppressed or if they can excite unintentional circuit resonances.

The nonlinear characteristics of the diode are also responsible for the dependence of in- and output impedance of the rectifier on power level. With a higher power level, bigger voltages across the diode are produced and more moderate parts of the characteristic curve are reached, effectively modulating the average r_D with power level.

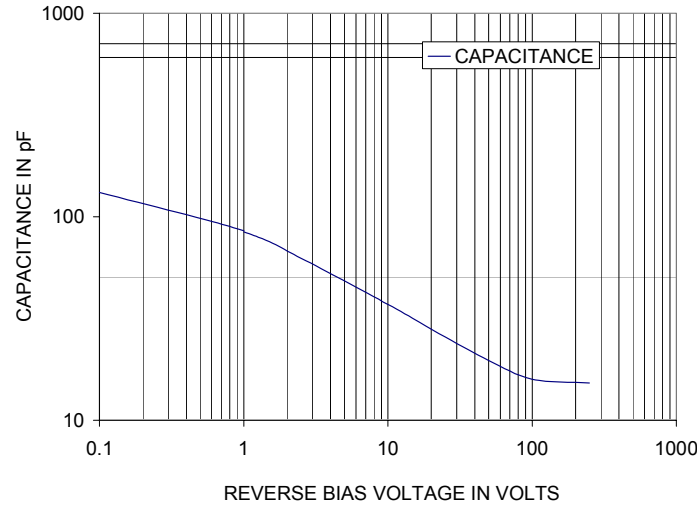


Figure 4.7: c_J over V_D for an IXYS GS250TI25110 Schottky diode

4.2.2 Junction capacitance c_J

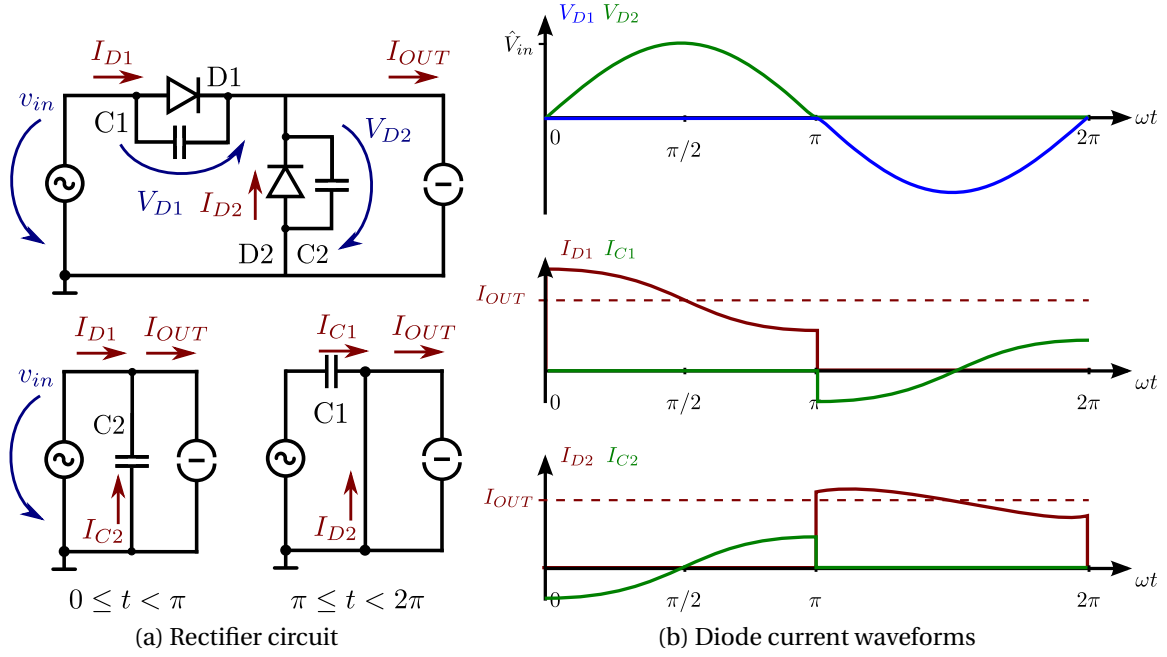
The diode junction suffers from a shunt capacity, whose value strongly depends on V_D in a nonlinear fashion. The origin and behavior of this parasitic is described in Chapter 5.1.2. The theoretical derivations in this chapter were only possible by assuming c_J to be constant. As long the stored energy stays the same, macroscopical operation is preserved to some degree (see Chapter 5.1.2). The waveforms with constant and voltage dependent c_J in Chapter 4.3 are very similar.

For a real diode the measured c_J is shown in a diagram in Figure 4.7. The junction capacitance is the main reason, why Schottky diodes for high power applications can not be used at RF frequencies. c_J has to be charged and discharged each cycle, resulting in huge imaginary currents.

Introducing c_J in the rectifier circuit, as shown in Figure 4.8a, changes the current waveform of the diodes considerably. The circuit operation can be explained in two half periods, looking at the current flowing in the first diode, which is the sum of I_{D1} and I_{C1} and shown in the second diagram in Figure 4.8b.

$0 \leq t < \pi$: The input voltage is positive, C1 and C2 are completely discharged. As v_{in} starts rising, D1 starts conducting and C1 can be ignored because it is shorted. The diode has to supply the constant output current I_{OUT} and additionally the charging current of C2. This results in an overlay of 2 waveforms for I_{D1} : A constant term arising from I_{OUT} and a cosine shaped term arising from C2 differentiating the input voltage. At $\omega t = \pi/2$, C2 has been charged to its maximum voltage and starts discharging again until it is at 0V at the end of the half cycle.

$$\begin{aligned}
 I_{D1} &= I_{OUT} + \frac{dv_{in}}{dt} \cdot C2 & I_{C1} &= 0 \\
 I_{D2} &= 0 & I_{C2} &= -\frac{dv_{in}}{dt} \cdot C2
 \end{aligned} \tag{4.21}$$

Figure 4.8: Analysis of a half wave rectifier, considering c_J

$\pi \leq t < 2\pi$: Exactly the same sequence occurs as in the first half cycle, only the component names D1 and C1 are exchanged with D2 and C2. That means C1 is now differentiating the input voltage and its charging current flows through I_{D2} .

$$\begin{aligned} I_{D1} &= 0 & I_{C1} &= -\frac{dv_{in}}{dt} \cdot C1 \\ I_{D2} &= I_{OUT} + \frac{dv_{in}}{dt} \cdot C1 & I_{C2} &= 0 \end{aligned} \quad (4.22)$$

As it can be seen, there is a certain symmetry present. The waveforms I_{D1} and I_{D2} are exactly the same but have a phase shift of one half cycle to each other.

In each half cycle, one of the parasitic C1 or C2 gets excited. Its current in response to the sinusoidal waveform can be seen 2 times each half cycle: through the conducting diode, overlaid by the output current and through the parasitic itself. This is important as this circuit behaviour stays the same when additional parasitic elements are introduced: the response waveform of the parasitics will still appear 2 times each half cycle although its waveform is considerably changed.

To prove this behaviour and show quantified values, a PSpice simulation has been done of the circuit. D1 and D2 are ideal switches, $R_L = 20\Omega$, $I_{out} = 1.55A$, $\hat{V}_{in} = 100V$, $V_{out} = 31.8V^2$. The currents trough the diodes and capacitors as well as the input voltage is shown in Figure 4.9.

4.2.3 Package inductance L_P

Adding a small package inductance results in a resonant LC circuit between L_P and c_J . It is formed by L_P and C2 in the first and by L_P and C1 in the second half cycle. One

²The relation $V_{out} = \hat{V}_{in}/\pi$ still holds true.

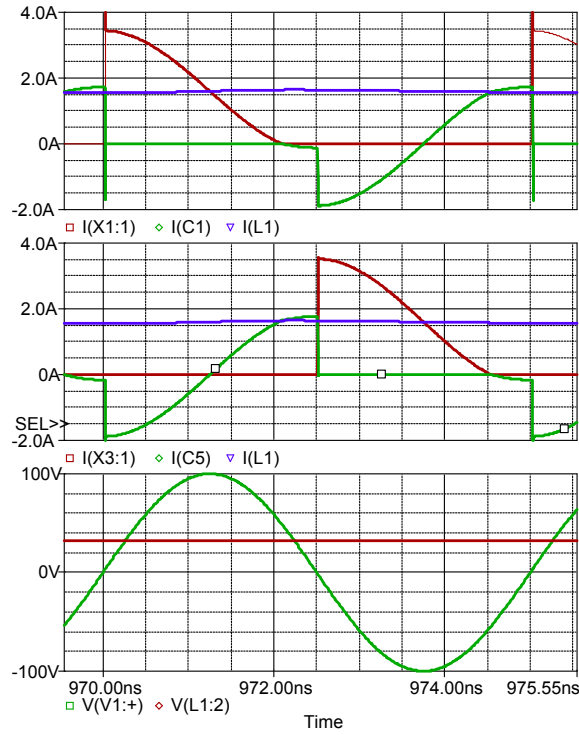
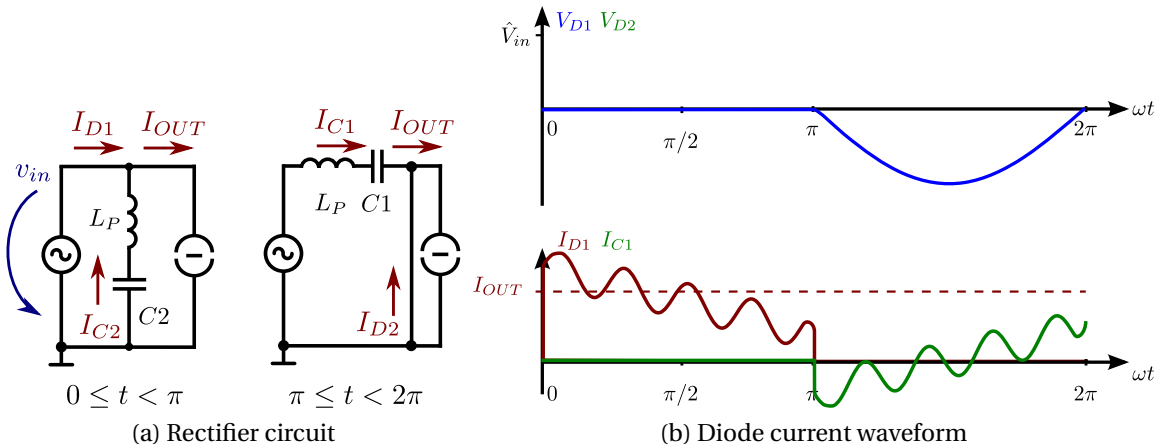


Figure 4.9: PSpice simulation of the rectifier

capacitance is always shorted by the conducting diode. The impulse response of the LC circuit is convoluted with the square wave shaped diode current from Chapter 4.1. The LC circuits impulse response is sinusoidal, its self resonant frequency is

$$f_{res} = \frac{1}{2\pi\sqrt{L_P c_J}} \quad (4.23)$$

Figure 4.10: Analysis of a half wave rectifier, considering c_J and L_P

The time constants of the LC circuits are much smaller than half a circuit period, so ringing happens with many cycles per period. This causes a high frequency current to flow in a loop through the other (conducting) diode and can be observed there too. The loops can be seen in Figure 4.10a.

This ringing is traditionally a problem in many non resonant rectifier and power converter circuits in the sub MHz range. Most of the passive components suffer from increased losses versus frequency. The high frequency AC current causes excessive heating of those components. Attempts to reduce ringing involve minimizing c_J and L_P , which in turn increases the ringing frequency and lowers the Q factor of the parasitic RLC circuit. The average junction capacitance is an inherent part of the diode. It depends on its power handling capability and is hardly under the engineers control. On the other hand, the loops which carry the high frequency currents can be minimized to a certain extent. Also the package can be optimized. After all there is a lower limit for L_P at about 1 nH which is simply dictated by component size.

A method often used in low frequency devices is to suppress the ringing by dissipative snubber circuits. Those devices have increasing losses with frequency. They dampen any higher frequencies and convert this energy to heat. The resulting loss decreases the efficiency of the overall rectifier. The energy stored in the parasitic LC circuit is lost. This can not be accepted in high frequency, high power applications, as diodes with big c_J have to be used. High efficiencies can not be reached without recovering the stored energy there.

4.3 Resonant operation

A solution to this problem is resonant operation [14]. The junction capacitance together with the package and a series inductance form a LC circuit, which limits the flow of high frequency currents. This results in a sinusoidal current flow through the rectifier. Harmonic radiation on the input is substantially reduced. The junction capacitance of the diodes is exploited as integral part of the circuit. The reverse voltage across the diodes rises slowly, allowing them to have some reverse recovery time (if PN devices are used).

In principle the ringing encountered in the previous section is exploited. An inductance, put in series with the diode lowers f_{res} , until there is no visible ringing anymore. The LC circuits oscillation is synchronized to the operating frequency.

4.3.1 Circuit analysis

Two equivalent circuits for each half cycle are given in Figure 4.11a. It is assumed that the diodes are ideal switches, the circuit is tuned, I_L is sinusoidal without any significant harmonic content and the output choke is big enough to not allow any significant change of I_{OUT} during one cycle.

During the first half cycle, D1 conducts and the upper equivalent circuit is valid. Looking at the currents of the circuit gives

$$I_L = I_{C2} + I_{OUT} \quad (4.24)$$

The current flowing into C2 during this half cycle is $I_{C2} = I_L - I_{OUT}$, meaning its waveform will be the input current offset by the output current. It can be seen in Figure 4.11b in the second diagram. At $t=0$ the capacitors current I_{C2} is zero, meaning all the input current flows into the output ($I_L = I_{OUT}$). As I_L increases above and decreases below I_{OUT} , a charging and discharging current flows into C2. At $t=0$ the voltage across

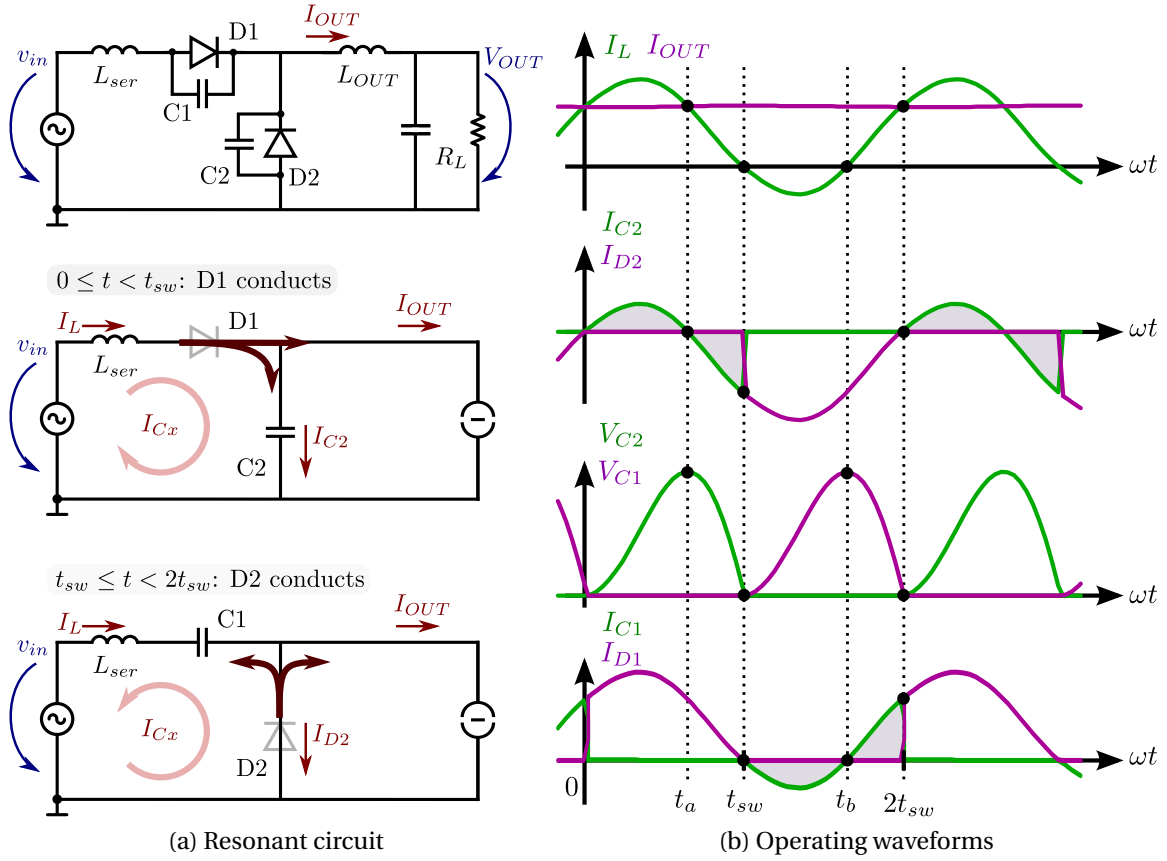


Figure 4.11: Equivalent circuits and waveforms of the resonant rectifier.

C_2 is zero. V_{C2} is the integral of I_{C2} (see Equation 4.25), so will be at a maximum at $t = t_a$, when $I_{C2} = 0$ or $I_L = I_{OUT}$.

$$V_{C2}(t) = \frac{1}{C_2} \int_{\tau=0}^{\tau=t} I_{C2}(\tau) d\tau \quad (4.25)$$

It is important to note that the average value of V_{C2} equals the output voltage V_{OUT} . This is the same relationship as derived in Equation 4.3 and will later be used to define the Q value of the LC circuit.

After $t = t_a$, until $t = t_{sw}$, the capacitor discharges again. The positive and negative areas under the curve (marked in gray) are the same, so the in- and outgoing charge equals and the capacitors voltage at $t = t_{sw}$ will be zero again. The component values have been chosen so that V_{C2} and I_L are zero at the switchover time $t = t_{sw}$. This way the on and off time of the diodes are symmetric, the duty cycle is 50 % and additional switching states – were both or neither diode conducts – are avoided. The current through C_2 equals $-I_{OUT}$ in that instance.

After the switchover point $t = t_{sw}$, the current through C_2 , commutates to D_2 as the diode starts conducting. This commutation happens almost instantly with very high current rise times. Fast current changes are usually a source of trouble in classical power converter circuits. However in this case, the commutation happens entirely inside the diodes package, the current changes over from the junction capacitance to the

actual junction and this change is not visible at the terminal pins of the device. Rather a smooth sinusoidal waveform, made up of the $I_{C2} + I_{D2}$ can be observed.

Now the second equivalent circuit becomes valid and the same charge and discharge cycle happens with C1. From $t = t_{sw}$ to $t = t_b$, C1 charges to its maximum voltage and discharges again completely as $t = 2t_{sw}$ is reached. As the area under the positive and negative part of the I_{C1} waveform is the same (marked in gray) the voltage is zero at the end of the cycle. The component values of C1 and L_{ser} have been chosen, so that the ringing transient – charging and discharging C1 once – takes exactly the time of one half cycle. At the end of the cycle ($t = 2t_{sw}$), the voltage of both capacitors is zero and the inductor current $I_L = I_{OUT}$. As this current commutates to D1, the initial conditions from $t = 0$ are once again reached and a new cycle can start from the beginning.

4.3.2 Characteristics of the rectifier

It can be seen, that L_{ser} forms a series LC tank circuit with C2 in the first, and with C1 in the second half cycle. Both are charged and discharged in smooth alternation by the ringing transient that happens when one switch opens. L_{ser} literally transports the junction capacitance's charge from one diode to the other. A purely reactive current loop is formed through V_{in} , L_{ser} and C2 in the first half and C1 in the second half cycle. The current is named I_{Cx} . To tune the rectifier into resonance, all the components from the current loop must form a LC circuit which is resonant on the input frequency. The condition for tuning is

$$f_{in} = \frac{1}{2\pi\sqrt{L_{ser}C_1}} \quad (4.26)$$

One very useful property of the resonant operation is, that any additional inductance in the I_{Cx} loop – originating from the diodes package pins or from circuit traces – can be absorbed by L_{ser} . The parasitic capacitance of the output choke L_{OUT} can be absorbed by C2 in the same fashion. This is possible as the DC side of the choke is shorted to ground for RF signals by the output capacitor. The rectifier is exploiting both, the parasitic capacitance³ and inductance of the diode.

During each cycle, some of the stored energy is transferred to the load resistor by I_{OUT} . This dampens the LC circuit and gives it a certain Q factor. Q is defined as the ratio of energy stored in the resonator by the power absorbed through I_{OUT} in each cycle. As the entire energy of the resonant tank is at a specific time throughout the cycle, stored in one of the 3 passive elements, the Q factor can be determined by looking at the real and imaginary power flowing through C2.

The real power, extracted from the LC circuit is given by

$$P_{real} = I_{OUT} \cdot V_{C2} \quad (4.27)$$

³It shall be noted that it is not advisable to put any additional, external capacitor across the diode as this will cause high frequency ringing with the diodes package inductance as the diode turns on and the fast current commutation happens [27]. A device with a bigger chip area should be used instead which also results naturally in a higher power capability.

The circulating, imaginary power in C2 can be calculated by

$$P_{imag} = \frac{V_{C2}^2}{\Im(Z_{C2})} = V_{C2}^2 \omega C2 \quad (4.28)$$

So the Q factor is

$$Q = \frac{P_{imag}}{P_{real}} = \frac{V_{C2}}{I_{OUT}} \omega C2 = \frac{\text{avg}(V_{C2})}{I_{OUT}} \omega C2 = \frac{V_{OUT}}{I_{OUT}} \omega C2 = R_L \omega C2 \quad (4.29)$$

The average of V_{C2} equals the output voltage, this is shown in Equation 4.3. Thus the resonant LC circuit is dampened by R_L which is effectively connected in parallel to C2 – the output inductor has no influence, as its average voltage over one cycle is zero. In fact, the equivalent circuit Figure 4.12 of the resonant tank can be derived from this assumption.

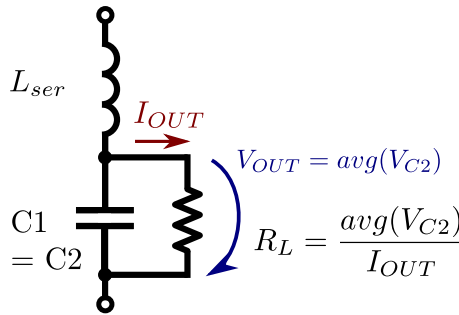


Figure 4.12: Equivalent circuit of the resonant tank.

Choosing a small R_L results in a low Q factor, increases I_{OUT} and lowers the reactive current I_{Cx} , resulting in more efficient power extraction, a higher input bandwidth but also a high harmonic content in the rectifiers waveforms. Higher Q operation is obtained by bigger load resistances, resulting in less harmonic content, but also more voltage and current stress on the rectifier and more resistive losses through big reactive currents. This behavior can be recognized in the simulations from Chapter 5.5.1.

The circuit operation is very closely related to that of a class E amplifier. Load and source on in- and output are exchanged, the switch is replaced by a diode. Exactly the same waveforms can be observed, they are only reversed in time [14, 27, 28]. The biggest difference is, that in a class E amplifier, the duty cycle is externally controlled by the mosfet's driving circuit. In the rectifier, the diodes turn on as soon as there is a positive voltage across them, making the diodes conduction duty cycle dependent on input voltage, load resistor and the L and C values [29]. The theoretical maximum power – for given diodes – is reached when both diodes conduct 50% of the time. For other duty cycles, there will be times where both or none of the diodes conducts.

One fundamental disadvantage of all class E like resonant circuits is their high voltage and current stress they put on the actual switch. Especially if the circuit uses high Q values and a single switch. As the rectifier of this work uses two diodes, it can handle double the power than a single diode circuit and compensates this disadvantage to some degree.

4.3.3 Input impedance

As the equivalent circuit from Figure 4.12 is valid for both half cycles⁴, it can be used to determine the rectifiers input impedance.

$$Z_{IN} = j\omega L_{ser} + \frac{R_L \cdot \frac{1}{j\omega C}}{R_L + \frac{1}{j\omega C}} = j \left[\omega L_{ser} - \frac{R_L \cdot Q}{Q^2 + 1} \right] + \frac{R_L}{Q^2 + 1} \quad (4.30)$$

The phase shift φ between input voltage and current is given by:

$$\tan(\varphi) = \frac{\Im(Z_{IN})}{\Re(Z_{IN})} = \omega \frac{L_{ser}}{R_L} (Q^2 + 1) - Q \quad (4.31)$$

If Equation 4.23 was used to design the rectifier (and the RLC circuits resonance frequency equals the operating frequency) – Equation 4.31 will show, that it does not have a purely real impedance on its input.

By setting the series inductance to a smaller or bigger value than Equation 4.23 predicts, the imaginary part of the input impedance of the rectifier can be compensated. The phase shift between input current and voltage can be eliminated, making the rectifier look like a resistor from its input. This simplifies the design of the impedance matching circuit in Chapter 4.4. The price for this is, that the diodes will not operate with 50% conduction cycle, slightly reducing its power handling capability from the theoretical maximum.

For the equivalent circuit, where the junction capacities are assumed to be constant, a real input impedance can be obtained by setting $\varphi = 0$ in Equation 4.31 and solving for L_{ser} . This gives Equation 4.32, a condition for the series inductance, to make the rectifier behave like a resistor.

$$L_{ser} = \frac{C \cdot R_L^2}{Q^2 + 1} \quad (4.32)$$

In the real circuit, this condition will not hold true anymore. The junction capacities are heavily dependent – in a nonlinear way – on the diodes instantaneous reverse voltage. Considering this leads to nonlinear differential equations that do not have a closed form solution [30]. To tune the circuit anyway, it can be simulated in PSpice. All the parasitics are considered in their full nonlinear form by the PSpice diode model. In a series of transient simulations, the optimum value of L_{ser} can be found, giving a purely real input impedance. This was done in Chapter 4.3.4.

4.3.4 Simulation

To show the advantages of resonant operation, the circuit has been simulated in PSpice, the schematic can be seen in Figure 4.13. The diode GS150TA25110 from IXYS is modelled, including its non linear junction capacitance and all other non idealities. The package and trace inductance is provided through Lp1 - Lp4. The circuit is simulated at 200 MHz at an output power level of around 200 W.

⁴In the real rectifier, the reactive current I_{Cx} flows alternately through C1 and C2 for each half cycle. As both have the same value and the same initial conditions at the switchover point, they are replaced by a single capacitor. This way the same reactive current can be observed. The DC power extraction is modelled by the resistor R_L . The resulting equivalent circuit is a resonant tank, without rectifying properties – but with the same electrical characteristics as the resonant rectifier.

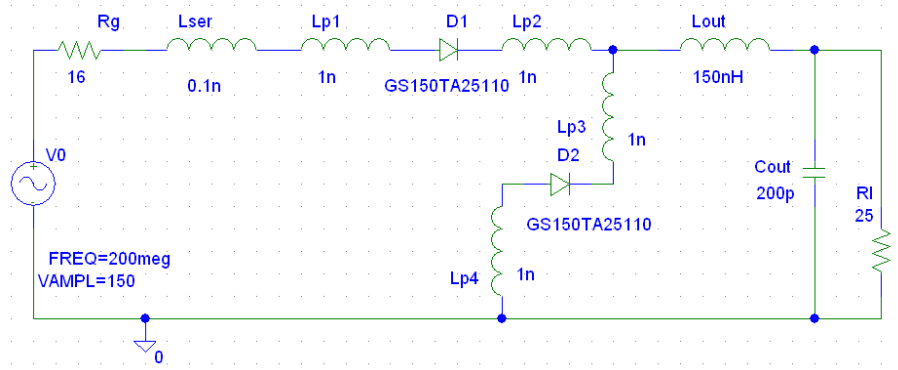


Figure 4.13: Schematic of the PSpice simulation of the resonant rectifier

In the first simulation, seen in Figure 4.14a, the top diagram shows the input voltage after the generator resistor (green) and the input current flowing through the source V0 (red). The current trough V0 consists of 2 parts.

- The constant DC output current, offsetting the current waveform. This is a problem for the realization of the circuit which is solved in Chapter 4.3.5.
- The reactive current I_{Cx} flowing in the loop formed by V0, R_G , L_{ser} , D1 and D2. For the non resonant case, this waveform contains a considerable amount of harmonics. (See also Figure 4.15)

Also the input voltage waveform has substantial harmonic content. There is a phase shift between current and voltage, meaning the input reactance of the rectifier is not zero. The voltage lags current, implying capacitive behavior, which is caused by the junction capacitance of the 2 diodes.

The second diagram shows the voltage across D2. The green waveform represents the measurement at the package pins, accounting for the package inductance. The red one represents the voltage directly at the diode junction. In the green waveform, ringing can clearly be observed each cycle. The difference between red and green waveform shows quite dramatically the voltage drop across the small package inductance and that measurements at terminals of the device can not be used to determine the internal state of the diode.

In the second simulation, Figure 4.14b, the series inductance $L_{ser} = 35nH$ has been added to the RF current loop. The high frequency currents are now restricted to flow by all inductors in this loop, resulting in a more sinusoidal waveform for input current and voltage. L_{ser} has been chosen to compensate the capacitive input reactance, resulting in zero phase shift between the input current and voltage waveform. Also the high frequency ringing across the diodes is less. The price for these improvements is, that the voltage stress on the active device has nearly doubled, showing the compromise that has to be gone with high Q resonant rectifier circuits.

To understand the behaviour of the circuit with a real diode, c_J is assumed to depend on the **average** voltage across the diodes and so directly on V_{OUT} . The parameters of the resonant circuit will change with output voltage, so with R_L and input power. Changing those values impacts conduction time of the diodes and the rectifiers complex input impedance. The dependency of all these values on R_L actually allows some degrees of freedom in circuit design and simplifies circuit operation:

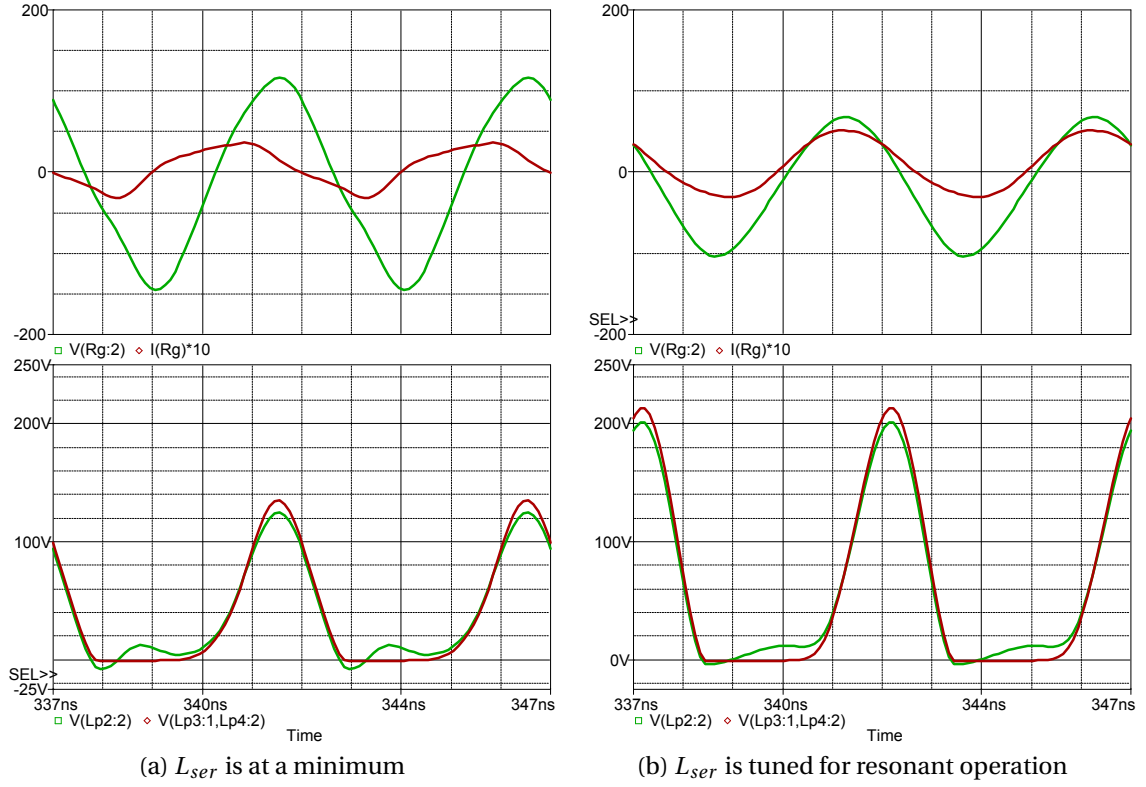


Figure 4.14: Resonant operation effectively reduces harmonic radiation on the rectifiers input but also causes more voltage stress for the active device.

- In the practical realization of the rectifier, R_L can be used to tune its input impedance to a purely real value. To explain this, all component values are assumed to be constant. If R_L is increased, several things happen. The first thing is the output voltage rises as the rectifier behaves like a current source. In Equation 4.3 it has been shown that the output voltage equals the average voltage across D2, which will increase too. More voltage across the diode means that the average value of c_f will be smaller (Equation 5.6). This has direct influence on the rectifiers LC circuit – which allows the resonant operation – and in the end changes the input impedance. This is exploited in the prototypes, as it is difficult to make the values of L_{ser} changeable.
- On the other hand, assuming the rectifier will be tuned by R_L , the series inductance L_{ser} will determine, what the optimum value for R_L will be. This allows to set the DC output resistance of the rectifier while designing the circuit.

The reason for this is again the variable c_f . To tune the rectifier containing a bigger L_{ser} , a smaller c_f is needed for the same resonant frequency (Equation 4.23). That means a bigger average voltage across the diode is needed, demanding a higher valued load resistor and resulting in a higher impedance operation (more voltage across, less current through the diodes).

This can be used to set R_L to an value, where the maximum voltage and current ratings of the diode are almost reached but never exceeded. This way the maximum power can be rectified with one particular diode.

It shall be noted that resonant operation is not a miracle solution to use arbitrarily large power diodes at RF frequencies. Using higher powered diodes with a bigger junction capacitance demands for a lower series inductance to keep f_{in} constant and still meet the resonant condition from Equation 4.23. As the series inductance can not be made arbitrarily small, this prohibits the use of very large power diodes with junction capacities in the nano farad range. The minimum realizable inductance from the prototypes of Chapter 7 – including circuit traces and package pins – showed to be 5 nH, setting an upper limit of $avg(c_J) = 130 pF$ to meet the resonance condition at 200 MHz.

The other limiting factor is the reverse voltage of the used diodes. Diodes with higher V_{rev} can usually handle more power for the same c_J (Chapter 5.4). Using such diodes results in a higher impedance operation of the circuit, the output voltage and so the voltage across the diodes will be bigger. Still, c_J has to be charged to the maximum diode voltage and back to zero in one cycle, which – for higher voltages – can only be done by higher currents. The energy to charge c_J is in principle not lost in a resonant circuit – the purely reactive charging current is provided by the series inductance. Problems are rather caused by resistive losses in the passive components and circuit traces, which increase proportional to the square of the flowing imaginary current. This places a limit on the use of high voltage diodes. Simulations showed that the reactive charging current will exceed the diodes current ratings easily. As it is not flowing through the actual semiconductor junction⁵, the datasheet ratings might be exceeded but R_S and at last the current rating of the bonding wires will place a limit on maximum charging current and thus on maximum usable diode reverse voltage.

4.3.5 Radiated harmonics

It has been shown in Chapter 4.2.1, that the nonlinear characteristic of the diode (describing an exponential characteristic) generates new harmonic frequencies. The diode behaves as an active current source, supplying rather than absorbing power on these frequencies. For an ideal half wave rectified current waveform in the time domain, the fundamental and every even harmonic is needed in the frequency domain.

It now depends on the peripherals of the diode, whether those harmonics are transferred to the input of the rectifier or not. It has been shown in Chapter 4.3 that a carefully sized inductance in the RF current loop can attenuate the high frequency components substantially and allow a beneficial resonant operation. The improvement can be seen clearly in Figure 4.15, showing the frequency components of the input current. Especially the second harmonic at 600 MHz is reduced and more energy is present on the fundamental.

To provide a path for the DC output current, a shunt LC tank circuit is added to the input. It prevents DC current to flow through the generator. This is important, as unnecessary power would be dissipated in the generator resistance. In fact the antennas

⁵The current through the diode can be split in one part, flowing through the junction when the device is forward biased and another part flowing through c_J when the diode is reverse biased. They can be seen as I_{C2} and I_{D2} in Figure 4.11b. I_{C2} causes power dissipation in R_S (referencing to the equivalent circuit of the diode in Figure 4.4) while I_{D2} causes dissipation in R_S and r_D . As the diode is for at least 50% of the time in reverse condition, less power is dissipated – compared to operating the device with a DC current. The maximum forward current ratings in a diodes datasheet are dictated by the maximum power dissipation with a DC current and so might be exceeded by a purely reactive current causing less dissipation.

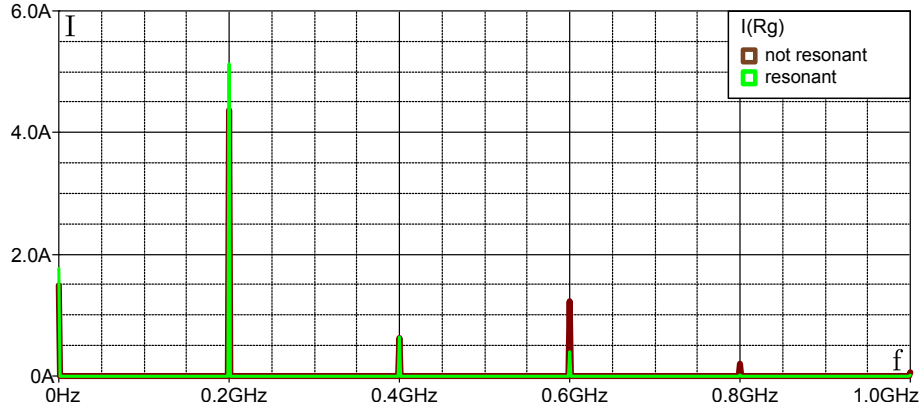


Figure 4.15: Frequency components of the input current.

of the coaxial power divider are not able to sustain a DC current at all. With the LC circuit, the line at $f=0$ disappears from the spectrum, no DC current flows through the source. The waveform of the input current gets centered around the horizontal axis.

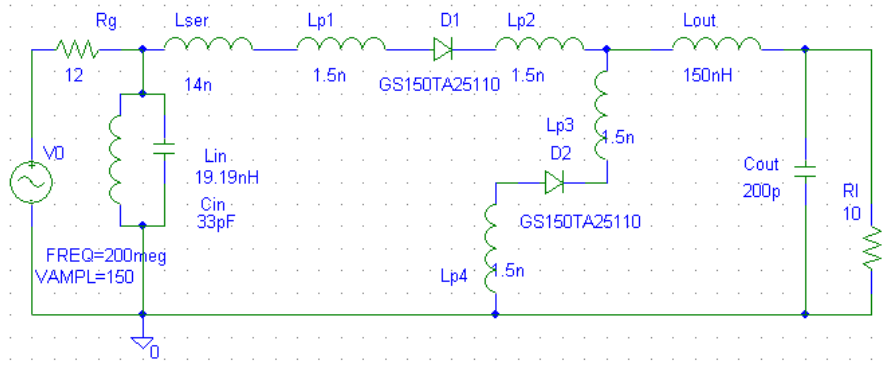


Figure 4.16: Rectifier with added LC circuit to form a DC path.

The LC circuit, together with the generator resistance of the input source, forms a parallel RLC circuit. It has the positive side effect of acting like an input filter, attenuating higher frequency harmonics. However, as the generator resistance is low ohmic, the filter will have a wide bandwidth. How much the filter reduces radiated harmonics, is investigated in the following section.

The RLC circuits impedance can be calculated analytically (Equation 4.33), it is plotted in Figure 4.17:

$$\frac{1}{Z} = \frac{1}{j\omega L} + j\omega C + \frac{1}{R} \quad Z = R + \frac{j\omega L}{1 - \omega^2 LC} \quad (4.33)$$

Its maximum impedance is at the resonant frequency $\omega_{res} = 1/(\sqrt{L_p C_j})$ and is defined by the parallel resistance. Below ω_{res} the inductor determines the circuits impedance by $Z = j\omega L$. Above ω_{res} the capacitor does by $Z = 1/(j\omega C)$.

Its resonance frequency is designed to equal the input frequency by Equation 4.23. This way it has a high impedance on the fundamental and low impedance on every other frequency, shown in Figure 4.17.

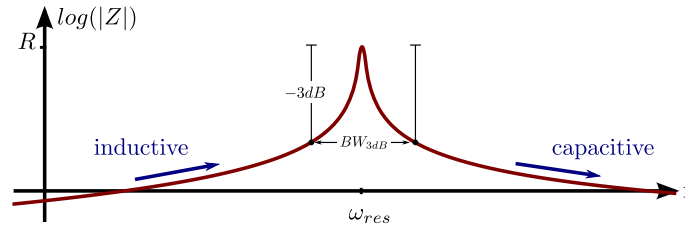


Figure 4.17: Impedance of an RLC circuit over frequency.

The parallel resistance determines the RLC circuits bandwidth. It is defined as the width of the impedance peak, between the points where the impedance is lowered 3 dB compared to the peak value. For a RLC circuit it can be calculated by taking Equation 4.33, setting it equal half the peak value ($Z = R/2$) and solving the equation for ω .

$$BW_{3dB} = \frac{1}{RC2\pi} \quad (4.34)$$

There is a degree of freedom in defining the L and C values for the same ω_{res} . Choosing a big C and small L results in a smaller bandwidth and good rejection of harmonics, as the capacitors reactance is smaller. It also results in big imaginary currents between L and C which has to be considered in circuit board layout, thick traces are needed to avoid excessive losses.

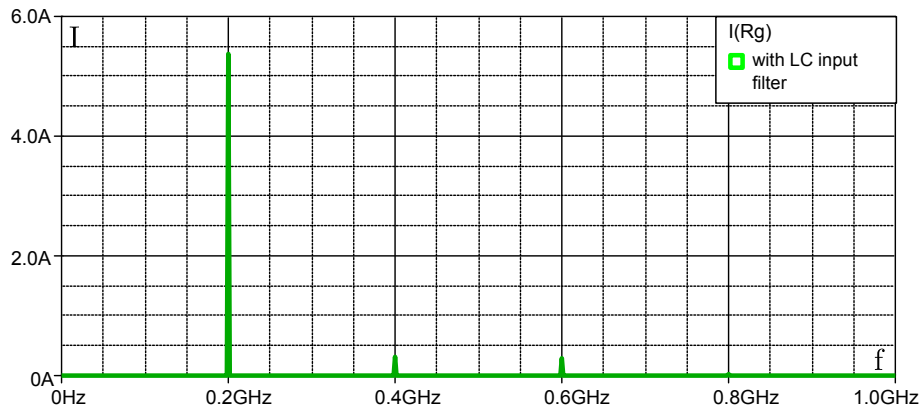


Figure 4.18: With added input filter.

The 3 dB bandwidth of the input filter, after Equation 4.34, with the chosen values is $BW_{3dB} = 402\text{MHz}$. This means the first harmonic at 400 MHz is reduced to half, the second harmonic to a quarter of its value. Higher frequency components are attenuated by the capacitor and harmonic radiation is reduced on the input of the rectifier. It can be seen in Figure 4.18 that the biggest harmonic is now 27 dB below the fundamental (4 %). If needed, this can be improved by more complex filters – but for this application the results are satisfactory and the simplicity of the LC tank is a big advantage for the actual realization of the rectifier.

4.4 Impedance match

In this section the input impedance of the rectifier is determined over input power level and solutions are found to match it to a 50Ω generator to obtain maximum power transfer.

The input impedance of the rectifier – operating in resonant mode – from Chapter 4.3 depends on the values of:

- c_J because the diodes junction capacitance is compensated by a series inductance, allowing the rectifier to operate in a resonance mode and having a real input impedance. If c_J is too big, the input impedance will become capacitive, if c_J is too small, it will become inductive.
- V_{DC} because the voltage across the diode changes its junction capacitance. The average diode voltage equals the DC output voltage.
- P_{in} because the output voltage changes with input power.
- R_L because the rectifiers output behaves like a DC current source and the output voltage changes with load resistance.
- L_{ser} because the series inductance is set to a specific value, so that the overall inductance in the RF current loop around the diodes compensates c_J at one specific power level. This value is chosen, so that the phase shift between input current and input voltage at the nominal input power becomes zero. Hence in this chapter the assumption $\Im\{Z_{IN}\} = 0$ can be made.

This means a perfect match can only be made at one specific operating point.

To determine the input impedance of a real rectifier, considering all parasitics, the circuit simulation program PSpice is used. The sinusoidal input current and voltage waveforms are displayed. The relation of their amplitudes reflects the magnitude R_{in} , their phase shift between each other reflects the phase angle Φ_{in} .

$$R_{in} = \hat{U}_{in(0)} / \hat{I}_{in(0)} \quad \Phi_{in} = \Phi(U_{in(0)}) - \Phi(I_{in(0)}) \quad Z_{in} = R_{in} e^{j\Phi_{in}} \quad (4.35)$$

A parametric simulation in PSpice, varying the input power from 1 W to 300 W shows the rectifiers $|Z_{in}|$ in Figure 4.19⁶.

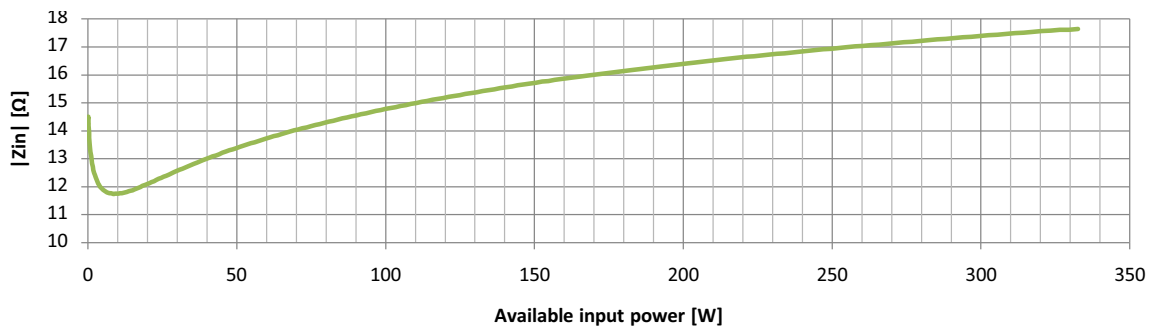


Figure 4.19: Input impedance over input power for the resonant half wave rectifier.

Only passive, non adaptive matching circuits are considered. The input impedance is determined at the nominal power 200 W, obtaining $Z_{in} = 16.5\Omega + 0j\Omega$. To match this to the 50Ω generator, several methods are possible:

⁶For technical reasons, the phase relation of fundamental in- and output current could not be extracted from PSpice. Nonetheless it can be seen in the practical measurements from Chapter 7

Magnetic transformers Magnetic transformers are common in the frequency range from 50 Hz up to 1 GHz. The upper frequency limit is dictated by the magnetic stray field, causing a high input impedance at high frequencies. Also capacitive parasitics pose a problem. A transformer was not used as the necessary core materials were not readily available in the laboratory.

Concentrated elements Resonant circuits consisting of 2 or more L and C elements can be used to transform impedance. They can be designed with a Smith chart very easily. Their disadvantage is that real inductors and capacitors are lossy and suffer from parasitics. This poses especially a problem at higher power levels and frequencies. They are also hard to get in arbitrary values.

TX-line transformers Transmission lines with a specific Z_0 and length, in series with the supply lines are used. The general idea is, that the reflections – created from the impedance step of the TX-line – cancel the reflections from the load by destructive interference. The source sees no reflected power whatsoever. This principle is also widely used in optical anti reflection coatings.

TX-lines can be used in different ways to facilitate an impedance match, this shall be examined further in the next section.

4.4.1 Quarter wave transformer

A series section of transmission line with an electrical length of $\lambda/4$ can be used for matching an arbitrary (real) load to any other (real) source impedance if its Z_0 is chosen right [31].

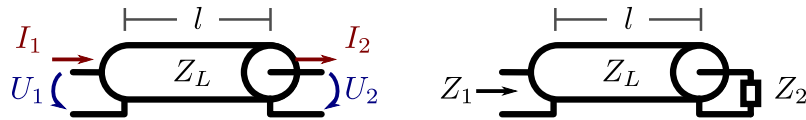


Figure 4.20: Voltages and currents on a TX-line.

In Figure 4.20 we see the voltages and currents at the end of a transmission line, they are related to each other by Equation 4.36. We assume the lines have no attenuation and are lossless.

$$U_1 = U_2 \cos(\beta l) + j I_2 Z_L \sin(\beta l) \quad (4.36)$$

$$I_1 = j \sin(\beta l) \frac{U_2}{Z_L} + \cos(\beta l) I_2$$

The phase constant β describes the spatial phase change along the transmission line in radians per meter. For a wave of given frequency or wavelength we get $\beta = \omega / c = 2\pi / \lambda$.

The relation between in and output impedance can be calculated by dividing the 2 equation for voltage and current from Equation 4.36 resulting in Equation 4.37. This result is very important, as it provides the fundamental rules for transforming impedances with transmission lines.

$$Z_1 = \frac{Z_2 + j Z_L \tan(\beta l)}{1 + j \frac{Z_2}{Z_L} \tan(\beta l)} \quad (4.37)$$

Now we consider a line with the length $l = \lambda/4$, inserting this in Equation 4.37 and solving the function in the limit, we get:

$$Z_1 = \frac{Z_L^2}{Z_2} \quad (4.38)$$

Thus we need a piece of transmission line with $Z_L = 28.5\Omega$ and $l_{el} = 0.375m$ to match the input impedance of 16.3Ω to the 50Ω source. The result can be seen in the Smith chart in Figure 4.21. The main problem with this simple approach is that a transmis-

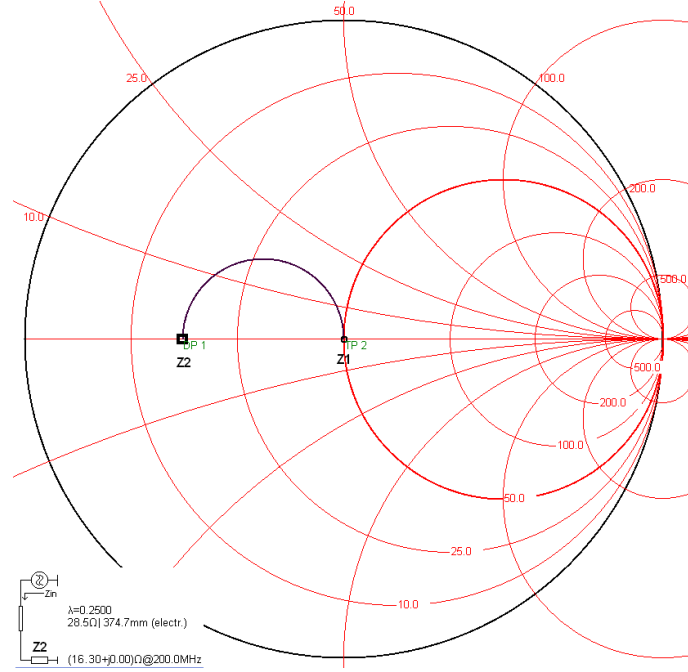


Figure 4.21: $\lambda/4$ impedance match.

sion line with an arbitrary value for Z_L is needed. For microstrip lines this is not a problem as the characteristic impedance can be adjusted with the width of the line. Coaxial TX-lines are generally only available in fixed standard values, meaning a compromise has to be taken and the next standard value used. This has been done for Prototype 1) in Chapter 7.2.1.

4.4.2 Two line transformer

To get around this problem, two lines in a series connection, with a given Z_l are used. Both lengths are variable, which gives 2 degrees of freedom, matching the real and imaginary part of any impedance.

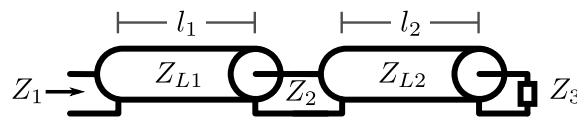


Figure 4.22: Two line transformer.

Again Equation 4.37 is used to calculate how the in and output impedance of the whole transformer relate to each other.

$$Z_1 = \frac{Z_2 + j Z_{L1} \tan(\beta l_1)}{1 + j \frac{Z_2}{Z_{L1}} \tan(\beta l_1)} \quad Z_2 = \frac{Z_3 + j Z_{L2} \tan(\beta l_2)}{1 + j \frac{Z_3}{Z_{L2}} \tan(\beta l_2)} \quad (4.39)$$

Both equations are combined and Z_2 is eliminated. The result is split in its imaginary and real part, resulting in 2 equations for the 2 unknowns l_1 and l_2 .

$$\begin{aligned} \tan(\beta l_1) \tan(\beta l_2) &= \frac{(Z_3 - Z_1)(Z_{L2} Z_{L1})}{Z_{L1}^2 Z_3 - Z_{L2}^2 Z_1} = a \\ \frac{\tan(\beta l_1)}{\tan(\beta l_2)} &= \frac{(Z_1 Z_3 - Z_{L2}^2) Z_{L1}}{Z_{L2}(Z_{L1}^2 - Z_1 Z_3)} = b \end{aligned} \quad (4.40)$$

Solving Equation 4.40 for L_1 and l_2 , we get two solutions for each line length from the quadratic functions. To avoid negative line lengths, the property of the tangent function – to be periodic in π – is used. The final formulas for the line length are shown in Equation 4.41.

$$\begin{aligned} l_1 &= \frac{\arctan(\sqrt{ab})}{\beta} & l'_1 &= \frac{\pi - \arctan(\sqrt{ab})}{\beta} \\ l_2 &= \frac{\arctan(\sqrt{a/b})}{\beta} & l'_2 &= \frac{\pi - \arctan(\sqrt{a/b})}{\beta} \end{aligned} \quad (4.41)$$

Those equations are used to design the improved impedance match for prototype 1) and prototype 2) in Chapter 7.2.3. For a given $Z_1 = 50\Omega$, $Z_3 = 16\Omega$, $Z_{L1} = 25\Omega$, $Z_{L2} = 50\Omega$ the two solutions $l_1 = 0.222m$, $l_2 = 0.0643m$, $l'_1 = 0.528m$, $l'_2 = 0.686m$ are obtained. They can be seen, visualized in a Smith Chart in Figure 4.23. In this case, the first solution uses far shorter lines, hence causes less attenuation, saves space and is realized in the prototypes.

4.5 Comparison of rectifier architectures

The resonant rectifier, analyzed in Chapter 4.1 showed fair characteristics and seems suitable for energy recovery applications. In this section the question shall be answered, whether other architectures would show superior properties. To find the most suitable rectifier architecture, six different types from literature are compared against each other in a simulation, shown in Figure 4.24. A consistent simulation procedure is used, which is described in Chapter 4.5.1. Different figures of merit are derived from the simulation data, which are explained in Chapter 4.5.2. Finally each rectifier architecture is shortly reviewed, in the end the best rectifier is chosen in Chapter 4.5.9.

4.5.1 Simulation environment

To make this possible, the architectures are designed for energy recovery in the SPS, operating at 200 MHz. For all simulation, the same diode model is used and the rectifier is optimized to convert the maximum power from RF to DC while staying in the safe area of operation of the diode.

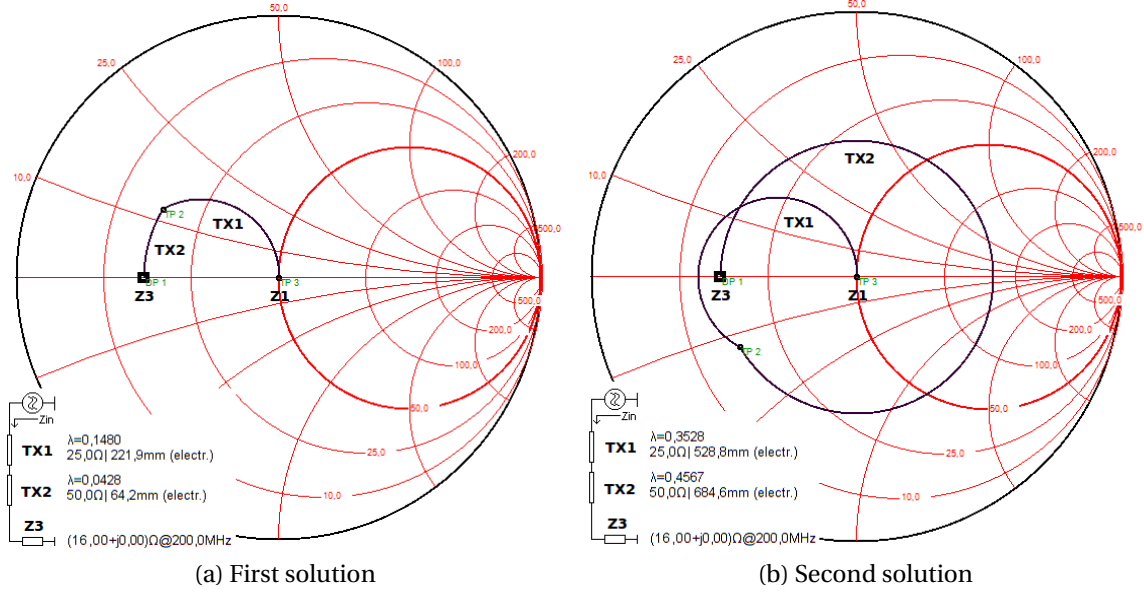


Figure 4.23: Matching the rectifiers impedance with two TX-lines.

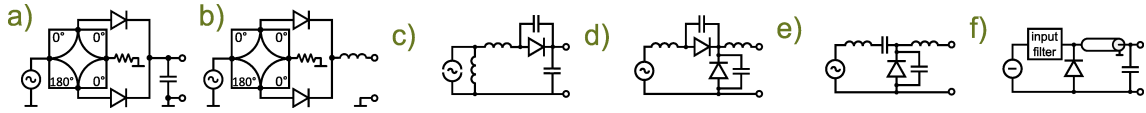


Figure 4.24: Rectifier architectures considered in the comparison.

The diode GS150TC25110 from IXYS is used. It is a high power GaAs diode in a special low inductance package for RF applications. Its main data is $I_{max} = 10A$, $V_{max} = 250V$, $c_J = 85pF(@1V)$. Its model is derived from the datasheet. The diagrams are curve fitted to extract parameters for the generic PSpice diode model. The software tool “PSpice model editor” is used for this purpose. The data for forward current versus forward voltage, junction capacitance versus reverse voltage, leakage current versus reverse voltage, and its reverse breakdown characteristics are taken from the datasheet and incorporated in the diode model. Its parameters are shown in Table 4.2. As it is a Schottky diode, the reverse recovery time by the means of minority carriers is zero through its intrinsic design. So the PSpice parameter “transit time” is set to a very small value.

Forward Current				Junction Capacitance	
IS	N	RS	IKF	CJ0	M
7.175 μA	1.4083	0.10183 Ω	0.68612A	118.1pF	0.40796
Rev. Leakage		Rev. Breakdown		Transit time	
VJ	ISR	NR	BV	IBV	TT
0.67629V	2.4358 μA	4.995	260.2V	2.519A	1e-016s

Table 4.2: PSpice diode model parameters for GS150TC25110.

After the modelling, crosschecks have been made, measuring the forward characteristic and junction capacitance in a simulation and comparing the diagrams with the datasheet, ensuring the diode model is realistic.

The parasitic package and lead inductances have been modeled by 3nH inductors on each diode pin. A typical schematic of a simulation can be seen in Figure 4.16.

For each rectifier simulation, 2 parameters are optimized in an unified procedure:

1. First the input voltage and generator resistance are set to the arbitrary values $V_0 = 200V$, $R_G = 25\Omega$, giving an available power of $P_{av} = 200W$ (Equation 4.42). This places the rectifier in an operational area of interest.

$$P_{av} = \frac{V_{0RMS}^2}{4R_G} \quad (4.42)$$

2. Now the load resistance R_L is adjusted to match the impedance of the diode. R_L and V_0 is varied and one cycle of the voltage and current of the diodes is plotted. The optimum R_L value is reached, when the maximum current and voltage ratings are approached, but not exceeded during regular operation.
3. Once R_L is determined, the reactive components (if existent) of the rectifier are tuned for resonant operation. Like in Chapter 4.3.5, input voltage and current waveforms are plotted. If there is no phase shift left, the optimum value is reached and the input impedance is purely resistive.
4. Now the source impedance is matched to the rectifiers input impedance⁷, without changing the input power level. This is important as the rectifiers behavior is heavily dependent on power level, so from its point of view, the same input power should be delivered. This is done by measuring the amplitude of input current and voltage on the fundamental frequency. The PSpice FFT function is used for that. Then the generator resistance and source voltage is set to: $R_G = \hat{U}_{in(0)} / \hat{I}_{in(0)}$ and $U_0 = 2\hat{U}_{in(0)}$.

Eventually steps 2, 3 and 4 have to be repeated until the optimum operating point is reached. After the circuit is tuned, a parametric transient simulation is run, varying the available input power from 1W to the maximum possible value. The figures of merit described in Chapter 4.5.2 are extracted.

4.5.2 Figures of merit

The most suitable rectifier fulfills all of the requirements listed in Chapter 3 in the best way. Following parameters are extracted from the simulations and used as a figure of merit to compare the rectifiers:

Max. DC output power How much output power can the rectifier supply, without exceeding the maximum current or voltage rating of the used diode(s). If one rectifier module can handle more power, the overall recovery system gets more economic, as the power divider gets less complex, less lossy and less pricey.

⁷This is possible, as impedance matching techniques are later used in the real circuit to transform the generators output impedance to any value (4.4)

RF/DC efficiency Defined as DC output power divided by available RF input power. Calculated in PSpice by Equation 4.43

$$\eta = P_{RL} / P_{av} \quad (4.43)$$

and visualized in a diagram over input power level.

T.H.D. Total Harmonic Distortion of the input current. Defined as the ratio of power on all overtones by the power on the fundamental. Gives a figure of how sinusoidal the input current is. A perfect undistorted sine is aspired, it has no overtones and a T.H.D. of 0%.

circuit complexity As several hundred rectifier modules are needed, its important for the circuit to be as simple as possible. The component count, production and tuning effort as well as reliability issues are evaluated to get a score from 1 (worst) to 6 (best).

exploited parasitics Does the rectifying architecture exploit the diode parasitics (c_j or L_p) as an integral part of the circuit? 0 means no, 1 means only a single parasitic is exploited, 2 means both are exploited.

number of diodes How many diodes are needed for the circuit. More diodes mean a higher price and a higher probability of device failure.

4.5.3 a) Voltage output full wave rectifier

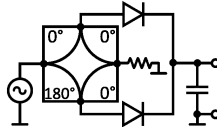


Figure 4.25: a) Voltage output full wave rectifier

For this rectifier, the simplest possible architecture – providing full wave rectification – has been chosen. It utilizes 2 diodes and a 180 degree phase shifter. The phase shift can be created by either a transformer with 2 inverted secondary windings (in the simulation), with a specific length of transmission line or with a 180 degree microwave hybrid (in the practical circuit). The circuit isn't tuned for resonance and no additional passive elements are added to allow a resonant operation. It is kept the simplest way possible. A full wave rectifier with four diodes and without transformer is not chosen, as in this configuration the effective junction capacitance is doubled, compared to a).

For the simulation, a transformer is used to get the 2 out of phase signals and to provide a DC path to the output. A capacitor in parallel to the primary is used to adjust for real input impedance and to dampen harmonic radiation.

The circuit is simulated as described in Chapter 4.5.1. It works best with a load resistance of $R_L = 6.1\Omega$. This value is too low for efficient DC power transmission, too much power will be dissipated in the wires resistance.

The current through the diodes and into the output capacitor is discontinuous, meaning it has large high frequency components and overtones. Those are only dampened

by the output capacitor, resulting in a critical high current, low inductance loop from the output capacitor to the diodes. This is hard to realize in an actual circuit and the cause of the low efficiency value of only 86 % and high T.H.D. of 13.7 %.

A 180 degree microwave hybrid is used in Prototype 0) to get the input voltage phase shift. The reason is that at 200 MHz, considerable engineering effort is needed to construct a transformer with low parasitic capacitance. The necessary materials were not readily available in the laboratory.

The DC output current will flow through the microwave hybrid. This has to be considered as the magnetic material in those devices can saturate, causing signal distortions. Depending on their internal construction, some hybrids do not provide a DC path at all.

It is interesting to note that the transformer version of the circuit effectively suppresses harmonics radiated toward the input. The 2 diodes generate the same harmonic spectrum. The diodes can be seen as a source for harmonic currents. Those flow into the transformer, cause a magnetic field of opposing polarity and cancel each other.

Also it has to be noted that a non resonant circuit generally bears less stress on the diodes voltage and current ratings. Higher input powers are possible at the cost of a reduced efficiency.

4.5.4 b) Current output full wave rectifier

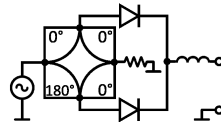


Figure 4.26: b) Current output full wave rectifier

The outcome of defusing the critical current loop of a) is this architecture. The output capacitor is replaced by a storage inductor. The circuit operation changes, now there is a continuous current flowing to the output at all times. The output current commutates smoothly between the 2 diodes, resulting in less T.H.D. of 6.9 % and higher efficiency of 92 %.

This architecture is based on the continuous current forward converter, which is found in power electronic literature.

4.5.5 c) Voltage output series resonant rectifier

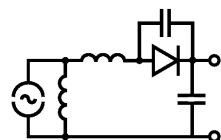


Figure 4.27: c) Voltage output series resonant rectifier

The series resonant rectifier with voltage output was taken from [18]. In this architecture a series inductance is placed on the diode. It is dimensioned carefully, so that the junction capacitance can resonate with it at the input frequency.

It is a common misconception to believe, that input current only flows during a small portion of the positive half cycle, like expected in a classic low frequency peak voltage rectifier with a single switch. As the diode with its c_J together with L_{ser} now rather behave like a resonant LC tank circuit, the input current of the negative cycle is stored in the passives and delivered to the output on the next positive cycle. With ideal components, 100 % efficiency can be reached in theory. In fact, the diodes conduction length is independent of the output ripple voltage [24] – unlike in a classical peak rectifier. Input current is sinusoidal and there are no peak charging currents occurring. This circuit has the advantage that both parasitics – c_J and L_P – of the diode are exploited for its operation, resulting in an excellent efficiency of 96.5 % in the simulation with the realistic diode model.

The shunt inductance across the input is needed to provide a path for the DC current. It can be introduced into the circuit without influencing its operation. Therefore the series inductor is split into two parts. One of it is converted by “series shunt transformation⁸” at the operating frequency to a parallel inductor.

A T - style input filter is used to keep the harmonics at acceptable levels. It consists of two 12 nH inductors and a shunt capacitor of 81.3 pF in between. The bandpass filter has been designed for lowest insertion loss at 200 MHz and has a 3 dB cutoff frequency at 240 MHz. This provides excellent rejection of harmonics but makes the circuit more complicated, especially as high quality capacitors and inductors at arbitrary values are needed.

For resonant operation always a compromise between high Q values – meaning high stress on the diode – or low Q values and high harmonic radiation has to be chosen. As the circuit uses only one diode, little power can be rectified while still maintaining a reasonable Q value. At the same time it is relatively complex, utilizing many components. This is the main disadvantage of the rectifier, as the application demands for simple high powered converters.

4.5.6 d) Current output series resonant rectifier

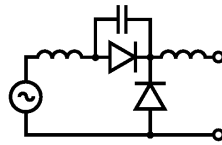


Figure 4.28: d) Current output series resonant rectifier

This is the continuous current version of c). The storage capacitor has been replaced by an inductor. A second diode has been added, to allow a continuous current flow through the circuit and – most importantly – to double the power rating, defusing the biggest disadvantage of c). The parasitics of both diodes are brought into resonance by adding a carefully dimensioned series inductance. The circuit operation has been analyzed in detail in Chapter 4.3.

⁸The idea of series shunt transformation is based on the fact, that any series combination of passive elements can be converted to a parallel combination with the same behaviour at one particular frequency. This can be shown mathematically.

The circuit shows very little harmonic distortion on its input waveforms by itself, a simple parallel LC tank circuit on the input is enough to reduce the radiated harmonics to an acceptable level of T.H.D. = 4.6 %.

The inductor at the output allows to connect a capacitor and that way to incorporate a second order output filter, providing an DC output with very little ripple.

This circuit suits the application very well, mainly because of its simplicity while still providing a very high output power of 321 W. The input filter is a simple LC resonant tank, having a minimum in component count and cost. The circuit needs two diodes but this poses no practical problems as many diode packages already contain two or more devices, sometimes already internally connected on the cathodes. This way, even the package inductance between the diodes is avoided.

4.5.7 e) Current output class E resonant rectifier

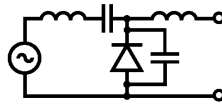


Figure 4.29: e) Current output class E resonant rectifier

This rectifier can be compared to d), leaving out the series diode but keeping the LC circuit. This is the 1:1 inverted version of an class E amplifier. Its design procedure and a theoretical analysis is shown in [28].

Its operation is closely related to d). It also provides a continuous current flowing through the circuit, but can be built with only one single diode. The finely tuned resonant operation puts the diode softly in its reverse blocking state, allowing the device to have some reverse recovery time. The distinguishing mark for class E operation is, that the voltage across the diode, as well as its derivative are zero at the instant of turn off. The energy stored in c_J is recovered by resonance, resulting in an excellent efficiency of 97 %. No input filter is needed, as the inherently sinusoidal operation of the rectifier generates no more than 3.0 % of T.H.D. at the input, which is acceptable.

Disadvantages of the circuit are its complex and restrictive design procedure. If the shunt capacity across the diode is entirely provided by c_J , there are not many degrees of freedom left to accomplish resonant operation and – for example – match the maximum voltage and current ratings of the diode. Also the circuit complexity is high in comparison to its output power. The package inductance of the diode is not exploited, it causes an RF ripple into the DC output and should be as small as possible [17].

4.5.8 f) Voltage output class F resonant rectifier

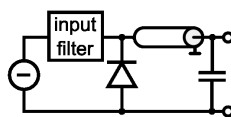


Figure 4.30: f) Voltage output class F resonant rectifier

This rectifier was used in some rectenna circuits at RF frequencies and has been taken from [22]. It consists of an input filter, used to block the radiation of harmonics to the input. One shunt mounted diode performs the rectification. The output filter accomplishes two tasks. It prevents ac components to appear across the load resistor and it provides a low impedance path for all even harmonic currents to flow. Even harmonics have the property of an zero average in one period of the fundamental, so they don't influence the average circuit behavior. An easy to realize output filter is a transmission line of $\lambda/4$ length, which is shorted by an capacitor on the end (this is also the DC output). Even harmonics ($2\omega, 4\omega$) are allowed to flow in a short high current loop through the diode and the transmission line. At odd harmonics ($\omega, 3\omega, 5\omega$) the transmission line presents a high impedance, restricting current flow. The DC output can pass without voltage drop. The circuits operation can be compared to that of a class F amplifier. Not considering any parasitics, the voltage across the diode will be shaped to a square wave, only containing the fundamental and odd numbered harmonics. The current through it will be shaped to a sinusoidal waveform, with one period clipped to zero, only containing the fundamental and even numbered harmonics.

Unfortunately none of the diodes parasitics is exploited in the circuit and introducing them in the simulation changes circuit operation considerably. Also the input filter is complicated to realize, needing at least 3 LC circuits which have to be tuned to the fundamental and harmonic frequencies. After all the circuit complexity is not in relation to the low power rating of only 129 W that could be reached with the rectifier.

4.5.9 Conclusion

In Figure 4.31 the magnitude⁹ of each rectifiers input impedance drawn over the available input power level can be seen.

The dependence on input power can be explained by the voltage dependent junction capacitance of the diode. At higher powers, there is a bigger voltage across the diode, severely reducing its c_j . It depends on the rectifiers circuit how the input impedance magnitude relates to this change. If the diode is shunt mounted the increased reactance also increases $|Z_{in}|$. An exception to this is circuit c), where the behaviour is inverted.

As an optimum match of $|Z_{in}|$ to the generator can only be achieved at a specific value of input impedance, a flat curve over a wide frequency range is appreciated. All but the circuit f) show a reasonable small variation of Z_{in} for matching at power levels $> 100W$.

In Figure 4.32 the RF/DC efficiency of the rectifiers is shown over input power level. The efficiency should be high over a large range of input powers. The absolute maximum power level the rectifier can handle is another important aspect. Clearly circuit d) shows very good efficiency values while also being one of the most powerful rectifiers. Circuit e) is the most efficient rectifier and also squeezes the most power out of one single diode.

In Table 4.33 all important properties of the considered architectures are summarized. The cell background color shows the suitability from green = best to yellow = worst.

⁹Unfortunately the imaginary part of z_{in} could not be extracted from the PSpice simulation. The complete complex impedance z_{in} is measured and visualized in a smith chart in Chapter 7 though.

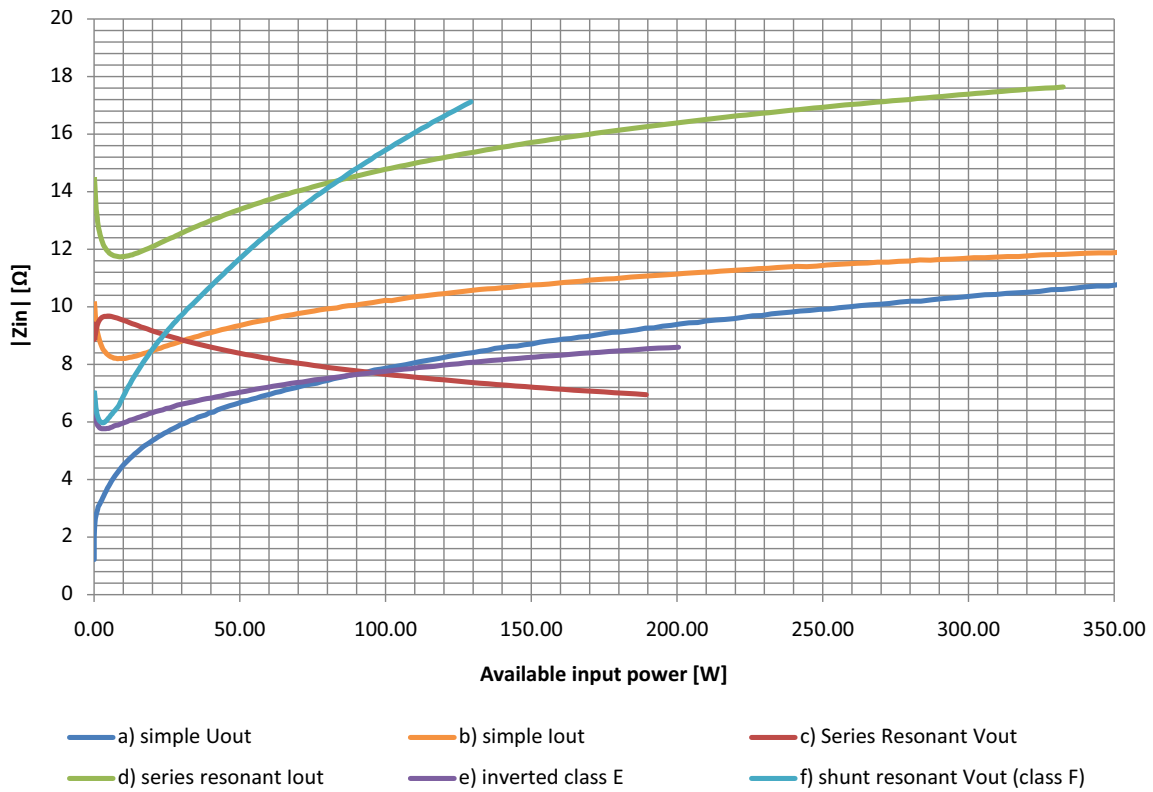


Figure 4.31: Magnitude of input impedance over input power.

The series resonant rectifier with current output d) has been chosen as most suitable circuit. It can handle high powers while being relatively simple in component count and easy to manufacture. Also its harmonic radiation is acceptable and it provides the second best values in efficiency. A prototype of this rectifier is designed, realized and measured for its performance in Chapter 7.

4.6 Synchronous rectifiers

In a synchronous rectifier externally controlled active switches like MOSFETs or HEMT transistors are used. The advantage is that modern mosfets can reach conduction resistances of $R_{DS,ON} = 1\text{ m}\Omega$ resulting in minimum forward losses. Just for comparison, a GS150TC25110 Schottky diode has a forward resistance of around $100\text{ m}\Omega$.

Another advantage is the controlling aspect. By varying the conduction angle, the output power can be controlled, making those devices very useful in applications like power managed computer CPUs, where a controllable DC voltage is needed. As we always want to recover as much power as possible, the advantage is not exploited in this application.

The 2 major disadvantages are:

- Active switches generally have a higher junction capacitance than diodes. It is this parasitic component that is limiting the RF/DC efficiency of the rectifier at RF frequencies. Using active switching devices doesn't provide any advantage, they have the same, usually even bigger $c_{J(D)}$ between drain and source than diodes. When using active switches, a second parasitic, the capacity $c_{J(G)}$ between gate and source has to be considered. This capacity is the reason, why

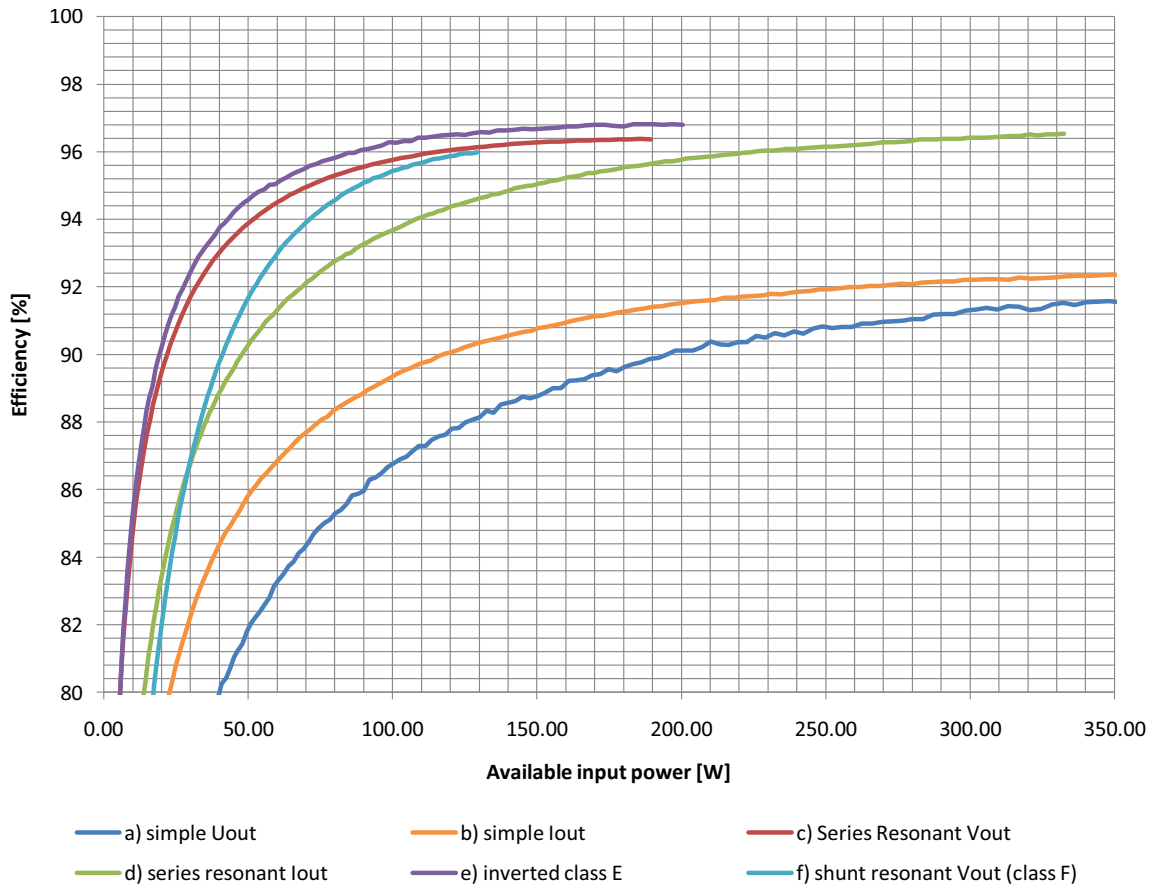
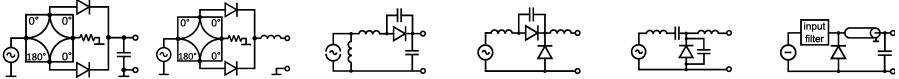


Figure 4.32: RF/DC efficiency over input power.

mosfets need a tremendous amount of gate drive power at RF frequencies in conventional power converters. There are some approaches, using resonant networks incorporating $c_{J(G)}$ and recovering this power [29] although they result in complicated circuit architectures.

- Every circuit architecture using active switches will be more complex, less reliable and more expensive than one using diodes. This is especially important, as some kind of resonance scheme can not be avoided in the control and power node for efficient operation at RF frequencies.

Alternatively, Schottky drain transistors might be considered. They can be operated in a diode like fashion by appropriately connecting the gate electrode. These devices are used in high power RF amplifiers at up to 2 GHz and promise a good performance for this application [32].



	a) simple Uout	b) simple Iout	c) Series res. Vout	d) series res. Iout	e) inverted class E	f) class F
Max. DC output power [W]	338	333	183	321	194.5	129
RF/DC efficiency [%]	86	92	96.5	96.5	97	96
T.H.D. On input [%]	13.700	6.956	0.004	0.092	3.029	0.345
circuit complexity [1-6]	2	2	3	3	5	6
exploited parasitics [0-2]	0	0	2	2	1	0
number of diodes [1-2]	2	2	1	2	1	1

Figure 4.33: Results of the rectifier comparison. Green = best, yellow = worst.

5. Semiconductors

The first diodes were made out of a silicon lump and a copper needle pointed into the crystal. This formed a metal semiconductor contact or Schottky diode. Diodes constructed like this were in use in the early days of radio transmission up to the 90s with an improved construction and hermetically sealed package. People found out that the electrical characteristics can be changed permanently by pulsing it with a high current. In fact part of the metal contact fused with the semiconductor, forming a N doped region and a proper PN junction. During the second world war, people found out that rusty razor blades can be used as rectifiers to build simple radio receivers. Construction methods have severely improved since then, especially Schottky diodes are in the focus of development and new findings in material research will makes those devices considerably improve the efficiency of present power converters.

5.1 PN junctions

The advantage of classic PN junction diodes is their high reverse voltage rating. Unfortunately they suffer from a certain reverse recovery time, which makes them unusable for RF frequencies.

5.1.1 Basic operation

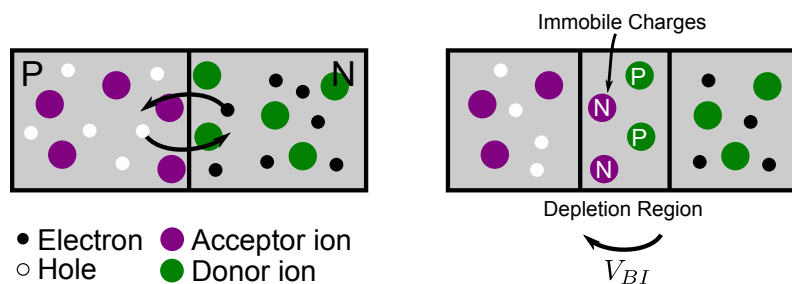


Figure 5.1: PN junction in static condition.

A basic PN junction without an external electric field is shown in Figure 5.1. The neutral (intrinsic) semiconductor material is doped with donor or acceptor ions, forming the N or P type silicon. Those are impurity's that are integrated into the silicons crystal lattice and provide free electrons (donors) or free holes (acceptors) as freely moving charge carriers. When a N and P type semiconductor is sandwiched¹ together, a diffusion process starts. The charge carriers around the junction are attracted to each other and recombine. The depletion region is formed, where all the electrons and holes neutralize each other, only the immobile donor and acceptor ions are left behind. The charge of those ions creates a voltage potential across the depletion region which is

called built in potential V_{BI} . It depends only on temperature and doping concentration of the silicon and is described by Equation 5.1 [33].

$$V_{BI} = V_T \ln \left(\frac{N_A N_D}{n_i^2} \right) \quad (5.1)$$

With $V_T = kT/e = 26mV$ for room temperature, $N_A = 10^{16}$ holes per cm^3 and $N_D = 10^{17}$ electrons per cm^3 for the doping concentration and $n_i = 1.5 \cdot 10^{10}$ which is the number of free electrons in $1cm^3$ of intrinsic silicon, the built in voltage $V_{BI} = 0.757V$ is calculated.

The actual spatial width of the depletion region can be calculated by Equation 5.2.

$$W_{dep} = \sqrt{\frac{2\epsilon_s}{e} \left(\frac{1}{N_A} + \frac{1}{N_D} \right) V_{BI}} \quad (5.2)$$

Where $\epsilon_s = 1.04 \cdot 10^{-12} F/cm$ is the electrical permittivity of silicon and $e = 1.602 \cdot 10^{-19} As$ the elementary charge of an electron. This gives a width of $W_{dep} = 329nm$ for this example.

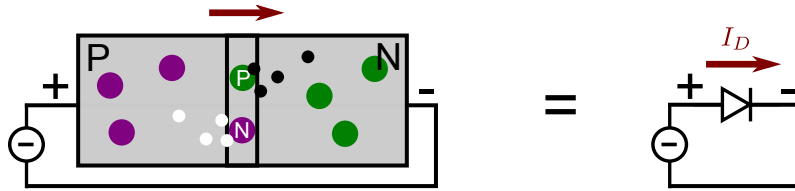


Figure 5.2: PN junction in forward condition.

What will happen when we apply an external voltage can be seen in Figure 5.2. The holes and the electrons are repelled by the external voltage and start moving in the direction of the junction. There they neutralize some of the immobile charges, effectively compressing the depletion region. As the external voltage rises and approaches V_{BI} , the depletion region gets smaller and vanishes entirely, allowing an increasing number of electrons and holes to recombine in the junction and an exponential rising current I_D to flow. The relation between voltage and current in forward condition is given by Equation 5.3, where V_F is the forward voltage across the diode, I_S the reverse current and n the ideality factor. It varies between 1 and 2 and depends on the fabrication process.

$$I_D = I_S \left(e^{\frac{V_F}{nV_T}} - 1 \right) \quad (5.3)$$

Reversing the external voltage, exactly the opposite happens. Electrons and holes get carried away from the junction, exposing more of the immobile charges. The depletion region gets bigger and the built in potential rises, effectively inhibiting current flow. To be precise, a small constant reverse current I_S is still flowing, caused by the minority carriers. These are charge carriers, randomly created by the non zero temperature

¹There must be one perfectly uniform crystal lattice, so literally putting them together is only a thought experiment. In real life this is done by taking one uniform intrinsic semiconductor material and creating the charged zones by a localized diffusion or ion implanting process.

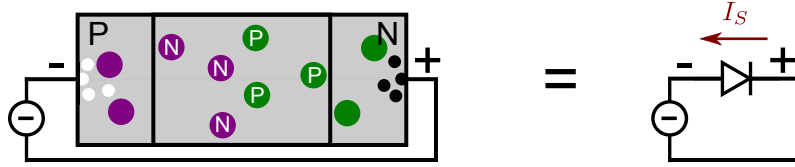


Figure 5.3: PN junction in reverse condition.

on the “wrong” side of the junction and recombine instantly. So I_S is only dependent on temperature and usually in the order of nA . However, once a certain reverse voltage across the diode is reached, electrons from the crystal lattice (not the free charge carriers) have enough energy to exit their covalent bonds. This causes an avalanche breakdown effect, rapidly conducting a very high reverse current. For Zener diodes this effect is cultivated and can be used for (noisy) voltage stabilization.

5.1.2 Origin of the junction capacitance

The depletion region can be seen as the dielectric in a parallel plate capacitor and thus a parasitic capacitance c_J is formed across the diodes junction. One unique property is that the depletion region changes its width depending on the applied reverse voltage, making the capacity voltage dependent in a nonlinear fashion.

The capacity of a parallel plate capacitor is defined by Equation 5.4, where A is the surface area of the plates and d is their distance to each other.

$$C = \frac{\epsilon \cdot A}{d} \quad (5.4)$$

Inserting the depletion regions width for d gives the physical formula for the junction capacitance of the diode with an reverse voltage. For the example above and a chip area of $5 \times 5 \text{ mm}$ $c_J = 79 \text{ pF}$ at 0.7 V is calculated.

$$c_J = \frac{\epsilon_s \cdot A}{W_{dep}} = \sqrt{\frac{\epsilon_s A^2 e}{2 \left(\frac{1}{N_A} + \frac{1}{N_D} \right) (V_{BI} - V_D)}} \quad (5.5)$$

More common in engineering work is the simplified descriptive formula used in PSpice [34]. It is shown in Equation 5.6 and can be parametrized to result in the same curves as Equation 5.5.

$$c_J = \frac{C_{J0}}{\left(1 - \frac{V_D}{V_{BI}} \right)^m} \quad (5.6)$$

The distribution of the doping ions in the junction is defined by m (linear $m=1$, step function $m=0$, hyperabrupt $m=-1$). C_{J0} is the capacity measured at $V_D = 0 \text{ V}$. The formula is valid for $V_D > 0$ too, but as the forward voltage is in the range of a few volts only, the capacity does not store much energy and is usually neglected. For a real diode the measured c_J is shown in a diagram in Figure 5.4.

The junction capacitance limits the maximum frequency, a Schottky diode can be used. High power devices have large chip areas, resulting in a c_J in the nano Farad range. It has to be charged and discharged each cycle, resulting in enormous imaginary cur-

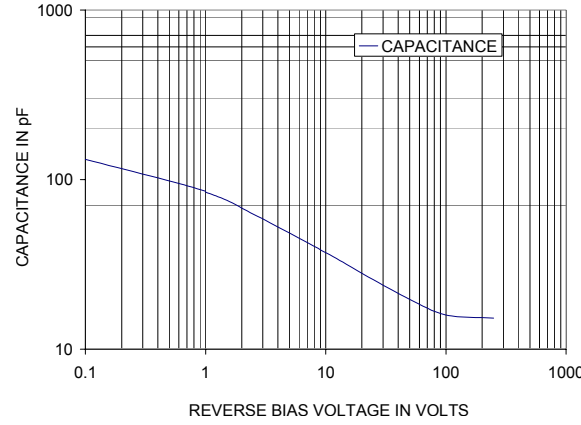


Figure 5.4: c_J over $-V_D$ for an IXYS GS250TI25110 diode

rents. In PN diodes, c_J can be neglected. Not because it is particularly small, but because the diffusion capacity is larger by an order of magnitude and makes PN diodes unsuitable for RF applications. The diffusion capacity is explained in Chapter 5.1.3.

It shall be noted that c_J can be exploited for a whole lot of practical applications. It is used as an electrical controllable capacitor – for example to shift the resonant frequency in LC circuits – with a static control voltage. Thus “varicaps” can be found in every radio receiver as part of an adaptive input filter.

Another interesting property is the generation of harmonic frequencies through the dependence of c_J on V_D . This can be shown mathematically by calculating the current through the capacitor [31]. The current through c_J is its change of electrical charge (Q) over time and can be calculated by derivation.

$$Q = C(t) \cdot V(t) \quad I = \frac{dQ}{dt} \quad (5.7)$$

$$I = \frac{dC}{dt} V(t) + \frac{dV}{dt} C(t) \quad (5.8)$$

$$I = \frac{dC}{dV} \frac{dV}{dt} V(t) + \frac{dV}{dt} C(t) \quad (5.9)$$

By assuming a sinusoidal excitation with DC offset, $V(t) = V_0 + V_1 \cos(\omega t)$, we get

$$I = \frac{dC(V)}{dV} \left[-V_1 V_0 \sin(\omega t) - \frac{V_1^2}{2} \omega \sin(2\omega t) \right] - \omega C(V) \sin(\omega t) V_1 \quad (5.10)$$

While the right part of the equation is the same as for constant capacitors, the left term shows interesting properties. Even without considering any particular nonlinear dependence between voltage and capacitance we can see, that the current now has a term on double the frequency of the input voltage.

This can be exploited in the use as an frequency doubler. Even amplifiers can be built using capacity diodes, exciting the diode with a sum of carrier and input signal. The input will be mixed to a higher frequency but also amplified. As capacitors don't shown

any noise whatsoever, amplifiers with exceptionally good noise figures can be built. This is used in helium cooled signal detectors for radio astronomy.

$$W_{var} = \int_T V(t)I(t)dt \quad W_{const} = \frac{1}{2}C_c V^2 \quad (5.11)$$

In the same fashion, the amount of stored energy can be calculated by Equation 5.11. W_{var} can be equated with the energy of a constant capacitor, W_{const} and the resulting equation solved for C_c . This value can be used in simplified simulations, as this constant capacitor stores the same energy and causes the same macroscopic operation of the circuit.

5.1.3 Origin of the reverse recovery time

The junction capacitance c_j is the dominating parasitic for PN or Schottky diodes when they are operated in reverse condition. If PN diodes are forward biased, there is an additional significant contribution in stored charge called the diffusion capacitance. It is the reason why the reverse recovery time t_s can be observed in PN diodes.

The typical current response of a reversed PN diode can be seen in Figure 5.5 as the red trace. An ideal diode would block the current immediately as soon as it reaches the horizontal axis. In a real diode the current is not altered at all for a certain time after it gets negative, the diode acts like a short circuit.

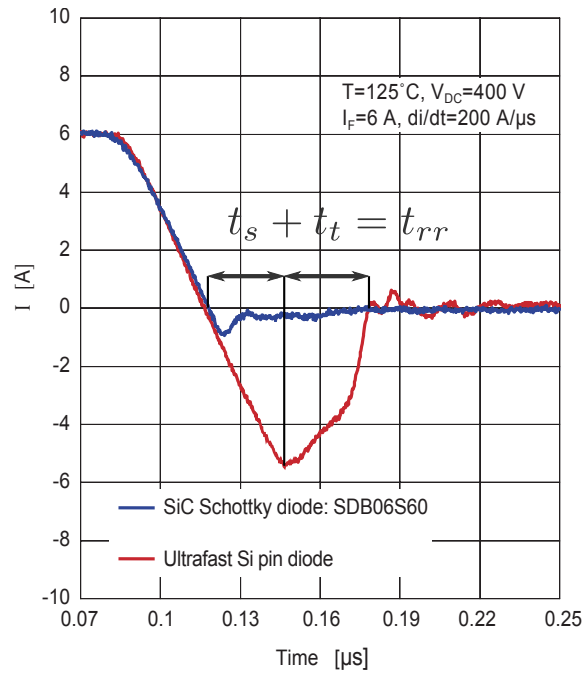


Figure 5.5: Typical reverse recovery characteristic of a diode. [35]

This can be explained by looking at the junction in forward condition (Figure 5.2). If it is conducting a current, all of the immobile charge carriers are neutralized by electrons and holes, that are pushed into the junction by the external voltage. A large number of minority carriers diffuse into and literally flood the respective – oppositely charged – regions (holes in the N region, electrons in the P region). They do not recombine with

their partners immediately but have a certain lifetime. The “spatial charge”, stored by minority carriers is given by

$$Q_{\text{dif}} = \tau \cdot I_{\text{fwd}} \quad (5.12)$$

where τ the statistical lifetime of the minority carriers. With a bigger I_{fwd} , the minority carrier density in the neutral region is higher. If they have a longer lifetime τ they can diffuse further into the N-type material. So the stored charge is proportional to both parameters.

When suddenly reversing the external voltage, the diode can not develop a depletion layer instantly. Rather the minority carriers have to be removed from the junction first. This happens mostly by random recombination in the junction. The time it takes for this process to complete is called the storage time t_s , it can be estimated by

$$t_s \approx \frac{Q_{\text{dif}}}{I_{\text{rev}}} = \tau \frac{I_{\text{fwd}}}{I_{\text{rev}}} \quad (5.13)$$

where I_{rev} is the average reverse current. t_s is dependent on the statistical lifetime of the carriers and the ratio of forward to reverse current. During t_s , the full reverse current flows but almost no voltage is measured across the diode, meaning that switching losses during this period do not occur in the diode itself but in its peripheral circuit elements. Only after the junction is clear and completely free of minority carriers, a depletion region can be established again, allowing a reverse voltage to be developed across the diode.

Now the junction capacitance still has to be charged, this happens during the transit time t_t . During that time, electrons and holes are stripped from the immobile charges, developing a voltage potential that is in balance with the external voltage. Charging c_j establishes the depletion region and puts the device into reverse state.

The overall time it takes to switch from forward to reverse state is called the reverse recovery time t_{rr} . For medium powered devices t_{rr} is generally bigger than 15 ns [36] and can reach up to 1 ms for high power diodes. As t_{rr} approaches the period of the RF signal, the device behaves as a current controlled resistor rather than a rectifier. The resistance for RF signals can be adjusted by the DC forward current. Indeed this principle is used in PIN diodes, which show small c_j but very big t_s values. They are used in variable RF attenuators and fast signal routing switches.

For rectifying purposes one is interested to minimize t_{rr} , it depends on τ which can be changed by different dopants, manufacturing processes, radiation treatments, etc. A common way to reduce t_{rr} is to introduce point defects in the crystal lattice. Those form spots where the minority carriers can rapidly recombine and so the junction is cleared faster. Point defects can be introduced by doping the junction with gold atoms, or exposing it to radiation. This way t_{rr} can be reduced by many orders of magnitude. The price for this is a higher reverse leakage current, a lower maximum reverse voltage and a slightly higher forward voltage. Especially the gold impurity's showed to be unstable and tend to age fast with increased temperature, weakening the rectifiers reverse recovery time with age.

5.2 Schottky junctions

PN diodes suffer from their reverse recovery time, which sets an upper limit on their usability in a high frequency rectifier. Schottky diodes get beneath that limitation, they operate by a different conduction process, have virtually no minority carriers in the junction and thus practically no reverse recovery time at all (blue trace in Figure 5.5). This makes them suitable for rectifiers at high frequencies.

5.2.1 Basic operation

A Schottky junction is formed by depositing metal on a polished n-type semiconductor surface. Its operation is best explained by an energy band diagram.

Electrons of a single atom can only exist in certain quantum states (or orbitals), depending on their kinetic energy. When bringing many atoms of a semiconductor together, the discrete states form continuous energy bands. Specifically for semiconductors and insulators at absolute zero temperature, all the bands, up to the valence band E_V , are completely occupied by electrons and do not contribute to the transport of charge carriers. Those electrons are fixed in place and form the crystal lattice by covalent bonds between the atoms. On the other hand, the conduction band E_C is completely empty. Between valence and conduction band is a band gap, where no electrons can exist. An electron needs to gain enough energy ($E_C - E_V$) to jump from the conduction to the valence band (for example through a temperature > 0 K), in order to move freely through the crystal lattice and act as a free charge carrier.

For metals, the band gap does not exist, valence and conduction bands overlap. Electrons can move freely at any temperature and do not need much external energy to cross a band gap, making metals excellent conductors. Insulators have a very wide band gap between valence and conduction band, resulting in a negligible amount of electrons reaching the conduction band, even at high temperatures. As the conduction band is empty and no free charge carriers are available, insulators have negligible conductivity.

The energy band diagrams of metal and a n-type semiconductor can be seen in Figure 5.6a, the x axis describes position, the y axis describes the energy of the electrons.

The Fermi level E_F is defined as the highest energy level which is still completely filled by electrons at a temperature of absolute zero. At higher temperatures, the quantum state with an energy of the Fermi level is filled to 50 % by electrons. E_F shows the statistical average energy of electrons in the band diagram. For semiconductors and insulators, the average energy level is between E_V and E_C , within the band gap (only statistically, as no electrons are allowed to exist there). The Fermi Level moves towards E_C with n type doping and towards E_V with p type doping. When a semiconductor device is in thermal equilibrium and there is no external voltage applied, the Fermi level will be constant throughout the whole junction.

The vacuum level E_0 is the energy of an electron outside the material and used as a reference for the other energy levels.

The work functions, Φ_M for metal and Φ_S for the semiconductor, define the voltage potential which is needed to free an electron from the Fermi level into vacuum. Multiplying the work functions by the elementary charge e gives the respective energies in electron volts and allows them to be drawn into the energy band diagram.

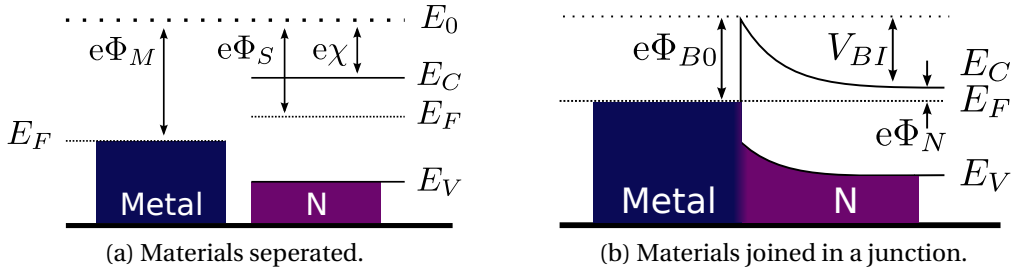


Figure 5.6: Energy Band Diagram for a metal semiconductor contact.

The electron affinity $e\chi$ defines how much energy is needed for an electron to jump from the conduction band (which is – unlike the Fermi level – independent of doping level) into vacuum.

When the two materials are brought together (Figure 5.6b), a system with a common, constant Fermi level is formed. Electrons with higher energy diffuse from the conduction band of the semiconductor into the metal. The immobile charges from the doping are exposed and a depletion region builds up in the semiconductor. This happens until the voltage potential is big enough to neutralize the diffusion force and thermal equilibrium is reached. A difference to PN junctions is, that the number of free electrons in metal is so large, that it is not possible to deplete it. The depletion region extends mostly to the semiconductor and only an insignificant distance into the metal. This can be compared to the situation with a PN junction, where the p side is doped tremendously. In fact, Equation 5.2 and Equation 5.5, describing W_{dep} and c_J , can be applied for a Schottky diode, if N_A is set to ∞ .

An important parameter from the band diagram is the barrier height, it is given by

$$\Phi_{B0} = \Phi_M - \chi \quad (5.14)$$

and defines the potential barrier, which electrons from the metal have to overcome, to flow into the semiconductor. In the idealized case its value is constant and independent of an external electric field.

Its counterpart on the semiconductor side is the built in voltage V_{BI}

$$\Phi_N = \frac{E_F - E_C}{e} \quad V_{BI} = \Phi_{B0} - \Phi_N \quad (5.15)$$

It determines the potential barrier, electrons from the conduction band of the semiconductor need to overcome, to get into the metal. Just like for a PN junction, V_{BI} varies with external voltage and doping level.

Once thermal equilibrium is reached, some higher energy electrons flow from metal to semiconductor and vice versa. This process is called thermionic emission (it would stop at 0 K). Both flows are balanced² and opposite to each other, so the net current is zero.

²Because both materials have the same Fermi level in thermal equilibrium, thus the electrons have to pass the same barrier height.

What happens, when an external voltage V_X is applied to the junction, can be seen in Figure 5.7.

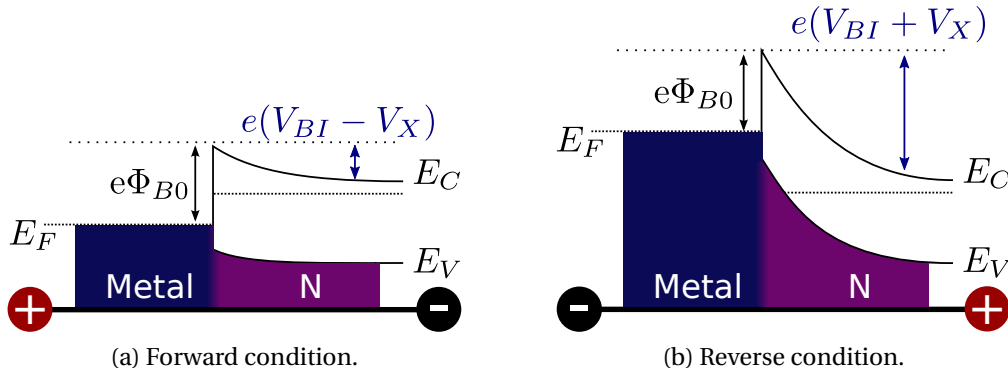


Figure 5.7: Metal semiconductor junction with external voltage.

Forward condition If V_X is positive at the metal, the electrostatic potential difference across the barrier will be reduced by its magnitude. Many electrons in the conduction band of the semiconductor have a high enough energy to move over the barrier into the metal. Therefore electron emission from semiconductor to metal is increased. The flow of electrons in the other direction is small because the barrier Φ_{B0} remains at its value from thermal equilibrium. This allows a net current to flow through the diode in forward condition.

Reverse condition If V_X is negative at the metal, the electrostatic potential difference across the barrier will be increased by its magnitude. Very few electrons can flow from semiconductor to metal, as they would need a lot of energy to overcome this barrier. However, some electrons are still emitted from metal to semiconductor, as Φ_{B0} remains at the same value from the equilibrium condition. This results in the diode's reverse bias saturation current I_S .

5.2.2 Comparison between PN and Schottky junctions

The most important difference of a Schottky, compared to a PN junction is, that no minority carriers (holes in this case) are needed for its operation. There is no recombination happening in the junction, electrons flow directly from the N type semiconductor into the metal. Thus the reverse recovery time t_{rr} for Schottky diodes is zero. No minority carriers have to be cleared and switching from forward to reverse operation happens almost instantly. Almost, because some current has to flow to establish the space charge region. This results in the junction capacitance c_J , the only major parasitic which limits high frequency operation. Small signal Schottky diodes with very small chip areas are used in signal detectors and mixers up to the tenths of GHz.

This process of current flow also allows higher current densities in the forward condition and allows Schottky diodes to have lower forward voltages $V_{BI} \approx 0.3V$ than their PN counterparts, which results in less conduction loss and makes them attractive as switches in power electronics or protection devices for transistors.

However, Schottky devices suffer from an inherently higher series resistance. Its cause is the need for a low doping concentration in the semiconductor material. At higher currents, the losses in Schottky diodes are dominated by the series resistance (see Chapter 4.2.1).

The magnitude of the reverse saturation current I_S is higher in Schottky diodes. It is caused by majority carriers which have enough energy to overcome the metal barrier height Φ_{B0} . Thus I_S is strongly temperature dependent. In PN diodes, reverse current results from minority carriers diffusing through the depletion layer and is small enough to be neglected.

If a metal conductor is attached to a very heavily doped n type semiconductor, the depletion region will be very small in reverse operation. In fact, if it is small enough, electrons can tunnel through the junction and an ohmic contact is formed. This is extensively used in microelectronics to build conductive traces and is important when building connections to semiconductors.

5.3 Wide Band Gap materials

Fortunately, the reverse leakage of Schottky diodes has been majorly improved in the recent years by research in Wide Band Gap (WBG) materials and band structure engineering³. State of the art Schottky devices are nowadays starting to be used in high voltage, high power applications, where they outperform the previously used PN diodes by orders of magnitude in switching speed.

Schottky diodes from materials as Gallium Arsenide, Gallium Nitride or Silicon Carbide can be operated at higher temperatures with way higher reverse voltages [37].

Property	Si	GaAs	6H-SiC	4H-SiC	GaN	Diamond
Bandgap, E_g (eV)	1.12	1.43	3.03	3.26	3.45	5.45
Dielectric constant, ϵ_r^a	11.9	13.1	9.66	10.1	9	5.5
Electric breakdown field, E_c (kV/cm)	300	400	2,500	2,200	2,000	10,000
Electron mobility, μ_n (cm ² /V s)	1,500	8,500	500 80	1,000	1,250	2,200
Hole mobility, μ_p (cm ² /V s)	600	400	101	115	850	850
Thermal conductivity, λ (W/cm K)	1.5	0.46	4.9	4.9	1.3	22
Saturated electron drift velocity, v_{sat} ($\times 10^7$ cm/s)	1	1	2	2	2.2	2.7

^a $\epsilon = \epsilon_r \cdot \epsilon_0$ where $\epsilon_0 = 8.85 \times 10^{-14}$ F/cm.

Figure 5.8: Physical characteristics of Si and the major WBG semiconductors (Source: [37])

The cause for this is that electrons need more external energy to jump from the valence to the conduction band, hence need to cross a wider *bandgap*. The external energy

³The idea of band structure engineering is to produce artificial materials, with certain electronic and optical properties. The materials characteristics are tailored to specific applications. Techniques used for this are alloying two or more semiconductors. Building heterostructures out of semiconductors with different band gaps (that constrain electrons in certain dimensions) or using the built in strain of mismatched crystal lattices.

can be given in the form of light (photons), radiation, a high voltage or by thermal movement. For GaN – compared to Si – over three times the external energy is needed to lift an electron to the conduction band (Figure 5.8), resulting in radiation hard, high temperature devices with a high breakdown voltage.

The higher breakdown voltage is a direct outcome of the wider band gap. This directly improves the three – on each other depending – parameters: Reverse voltage, forward current and junction capacitance. Basically it allows to use higher doping intensities, resulting in a lower barrier height for the same reverse voltage ratings. The “thinner” junctions have the same reverse leakage but less forward voltage losses. Also wbg materials show higher saturated drift velocities, which are important for high frequency operation of bipolar transistors and an important factor for determining the reverse recovery time of PN diodes.

Another very important factor – especially for this application – is the junction capacitance. It depends proportionally on the square root of the materials dielectric constant $\sqrt{\epsilon}$, the reverse voltage $\sqrt{1/V_D}$ and the doping intensity $\sqrt{N_D}$ (Equation 5.5). It also depends proportionally on the chip area A , which is designed by the diode manufacturer to accomplish a target forward voltage drop at the maximum rated forward current [38]. With wbg materials, smaller area chips with the same forward voltage thus same forward current rating – but less c_J can be designed. Also they are generally operated at higher voltages, further reducing the effective c_J during operation.

As some wbg materials show a higher thermal conductivity than silicon, it makes them more suitable for high power pulsed and CW applications. More thermal energy can be dissipated in a given amount of time. Together with the higher operating temperatures, excellent high power devices with higher power dissipation ratings, for a given package, can be made.

The main disadvantage of these materials is their difficulty in manufacturing and thus high price of the devices. Looking for example at SiC. Its melting temperature is higher than 2200 °C, requiring a vapor deposition process to grow the material. This is 50 times slower than manufacturing pure silicon by pulling and makes the material very expensive. Also different lattice structures – called polytypes – can grow in the crystal. This is sometimes denoted by a prefix in the name like 3C-SiC or 6H-SiC. Which polytypes will grow is very hard to control, as not all of the influencing parameters of the process are understood yet [39]. Sometimes so called micropipes (Figure 5.9) form in the material during the growing process, those are hollow channels with a diameter in the μm range that run through the crystal lattice. What micropipes are, why they form and exist at all is not yet clear either. Presently a lot of research is done on those phenomena. The formation of micropipes and non uniform polytypes are the main problems while growing SiC to date [39]. However, the semiconductor foundries will further refine their manufacturing process, to obtain higher quality and larger diameter wafers. The price of the material is expected to drop, as the demand in industry increases.

Another application specific disadvantage is, that WBG devices operate naturally at higher voltages. Only in this way, advantages from a higher electric breakdown field strength can be gained. In the resonant rectifier, low voltage and high current devices are preferred as they keep the reactive currents low (Chapter 4.3).

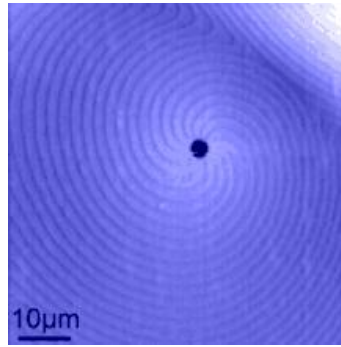


Figure 5.9: Micropipe which grew in SiC material. Source: [39]

5.4 Comparison of available diodes

To choose the most suitable diode for the rectifier, 14 different devices from different manufacturers and with different materials have been compared. The data has been taken from their datasheets. As the junction capacitance is voltage dependent, the value at 1 V reverse voltage has been taken (C_{j1}). Two figures of merit are derived from the diode's data. “Pmax” is the maximum forward current (I_f) times the maximum reverse voltage (V_r) and gives a hint of how powerful the diode is. P_{max}/C_j is the power normalized to the junction capacitance and the crucial factor in the decision which diode to choose.

Name	Material	V_r [V]	I_f [A]	C_{j1} [pF]	P_{max} [W]	P_{max}/C_j
BeMiTec custom	GaAs	70	1.8	4	126	31.5
GS150TC25110	GaAs	250	10	82	2500	30.5
IDD04SG60C	SiC	600	4	80	2400	30.0
IDD10SG60C	SiC	600	10	260	6000	23.1
C3D03060A	SiC	600	3	90	1800	20.0
DGS20-018A	GaAs	180	23	210	4140	19.7
DGS 3-018AS	GaAs	180	7	70	1260	18.0
STPSC406D	SiC	600	4	137	2400	17.5
STPSC406B	SiC	600	4	137	2400	17.5
C3D08060	SiC	600	8	300	4800	16.0
C3D10060A	SiC	600	10	500	6000	12.0
CSD01060	SiC	600	1	70	600	8.6
PMEG3020EH	Si	30	2	60	60	1.0
BAT160	Si	60	1	80	60	0.8

Table 5.1: Comparison of currently available Schottky diodes.

Diodes with a high P_{max}/C_j , a high absolute power and a low forward voltage⁴ are preferred. This is why the GS150TC25110 device from IXYS appeared to be most suitable after Table 5.1. It has the further advantage, that the diode is packaged in a way, perfectly suited for a series resonant rectifier architecture. The package contains 3 diodes with a common cathode and has special, very wide, metal sheet connectors (Figure 5.10a). Both are measures to cut the package inductivity L_p to a minimum.

To evaluate the device in a resonant rectifier, its PSpice model was derived from

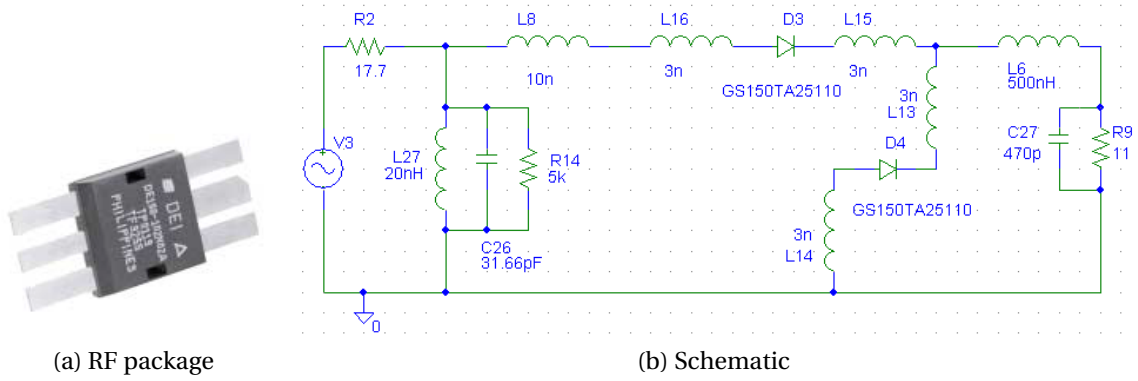


Figure 5.10: Simulation of the series resonant rectifier with IXYS GS150TC25110 diodes.

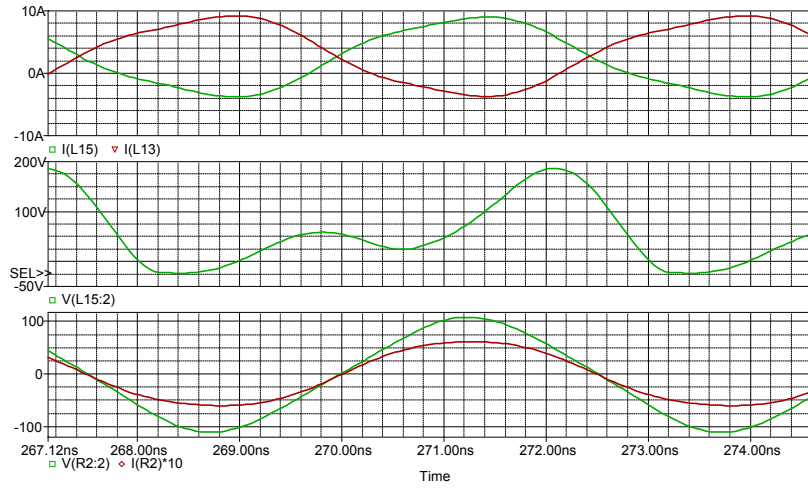


Figure 5.11: Results of the simulation with GS150TC25110 diodes. Top to bottom: Current through Diodes, voltage across the package pins of D10, input voltage and current*10 of the rectifier.

the datasheet and additional measurements. The series resonant current output rectifier was designed as described in Chapter 4.5.1. For an input power of $P_{in} = 333W$, $P_{out} = 320W$ DC output was produced at the load resistor. Efficiency with ideal passive elements was $\eta = 96\%$. The diodes are operating at lower voltages, so the average c_f is higher than with a IDD04SG60C diode. This means only a small series inductance $L_{res} = 10nH$ can be placed in the circuit to tune it to resonance. Care has to be taken when building the circuit as this doesn't allow much headroom for parasitic trace inductance's and similar. In fact, a too big inductance in the RF current loop was the reason prototype 2 didn't perform as expected. Also high frequency currents are less damped by the low series inductance, resulting in current and voltage waveforms with higher harmonic content, shown in Figure 5.11.

⁴Because higher voltage devices result in extensive reactive currents in resonant operation. See Chapter 4.3

5.5 Alternative devices

5.5.1 IDD04SG60C

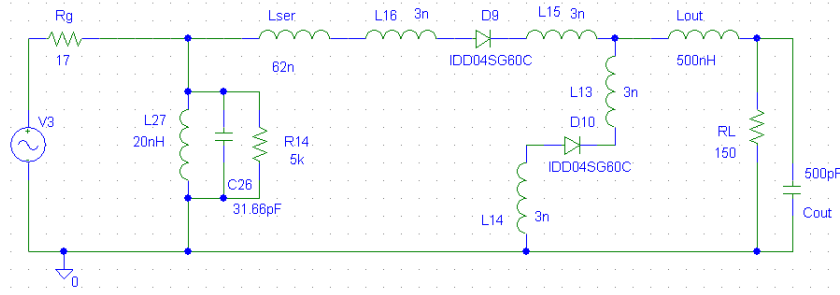


Figure 5.12: Series resonant rectifier with IDD04SG60C diodes.

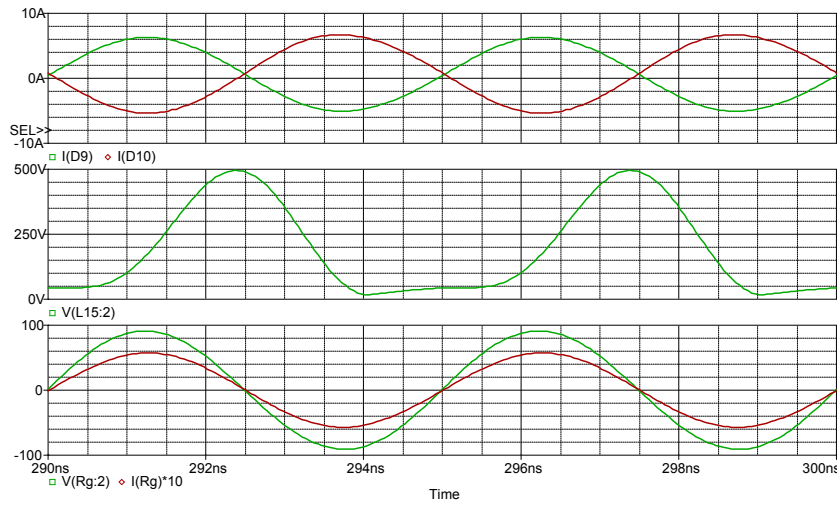


Figure 5.13: Results of the simulation with IDD04SG60C diodes. Top to bottom: Current through Diodes, voltage across the package pins of D10, input voltage and current*10 of the rectifier.

The device IDD04SG60C from Infineon was evaluated with the rectifier architecture of choice in a PSpice simulation. It is a SiC device in a common DPAK package. The device model was obtained from the manufacturers website. The rectifier was designed as described in Chapter 4.5.1, its power ratings are $P_{in} = 260W$, $P_{out} = 247W$, its efficiency with ideal passive components is $\eta = 95\%$. The diode is a higher voltage device rated at $V_r = 600V$, so for efficient operation a rather high ohmic DC load is needed ($R_L = 150\Omega$). The high voltage operation results in high reactive currents between the diode's c_J and the series inductance (Figure 5.15 top). As the reactive currents cause less power dissipation in a diode, they are allowed to slightly exceed the device's ratings. Input voltage and current were brought in phase (Figure 5.15 bottom) by setting $L_{ser} = 62nH$. The big series inductance corresponds to a very low average c_J of the diodes, originating from the high voltage operation. It leaves a lot of headroom relating to parasitic inductance of the circuit board traces and more effectively filters out harmonic components. Indeed all of the waveforms in Figure 5.15 look very smooth and free of high frequency components.

The advantages of the WBG material are effectively used in the rectifier. The diode has a forward voltage drop of only 2V at 4A which is negligible if the rectifier is utilizing the

full 600V reverse voltage. The high operating voltage minimizes the junction capacitance, allowing a lot of inductivity to be placed in the RF current loop. This also leaves headroom to meet the resonant condition at higher frequencies. The high operating voltage also causes large reactive currents which will cause ohmic losses in a real circuit and will exceed the datasheet's forward current rating of the diode. This might be legit but should be reviewed with the manufacturer in detail or investigated in an experimental setup. In fact by allowing more reactive current to flow, the output power could be further increased.

As it was not possible to get hold of sample devices⁵, a prototype was not constructed. Nonetheless this diode is one of the most promising ones for an energy recovery application at 200 MHz and should be considered in future work on this subject.

5.5.2 BeMiTec custom diode

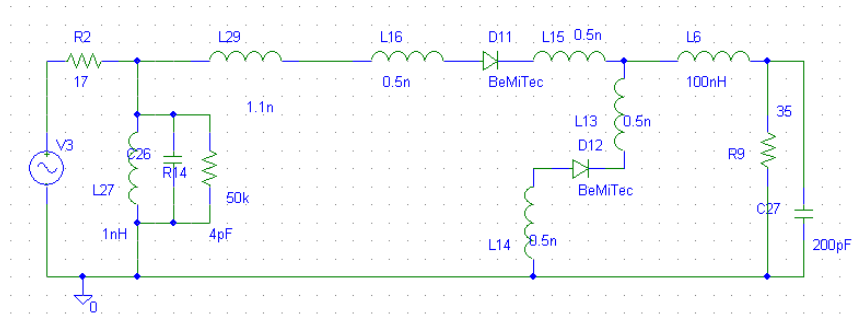


Figure 5.14: Series resonant rectifier with BeMiTec diodes.

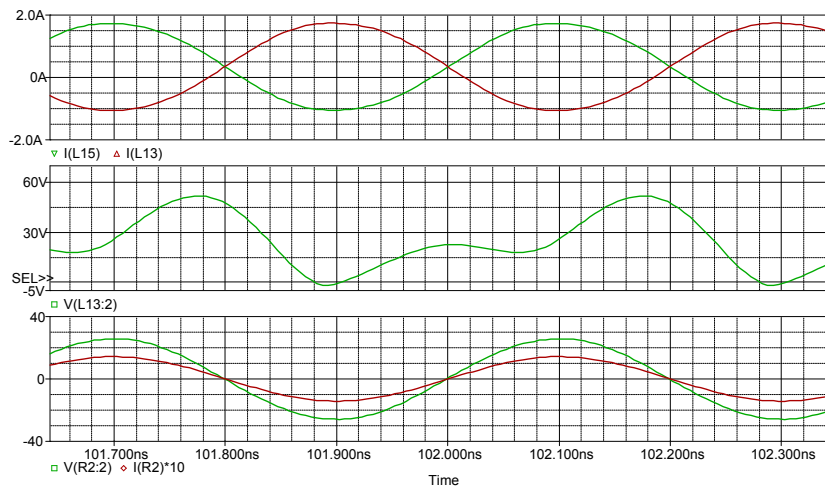


Figure 5.15: Results of the simulation with BeMiTec diodes. Top to bottom: Current through Diodes, voltage across the package pins of D10, input voltage and current*10 of the rectifier.

In Table 5.1, the device named “BeMiTec custom” has the best Power / Capacity rating. It is a GaAs Schottky diode, not yet commercially available, obtained through the “Ferdinand-Braun-Institut für Höchstfrequenztechnik” in Berlin [32]. The diodes were

⁵Infineon started production of those novel devices in 2009 and the present demand is very high, especially in the industry of automotive power converters for hybrid cars.

developed as blocking diodes for integrated switch-mode RF amplifier circuits and provide a unique combination of power and speed. A PSpice model has been created by data already available [40] and additional measurements⁶. All diode parasitics are considered. As this diode is tailored for higher frequency operation than 200 MHz, it was simulated and optimized for a microwave rectifier circuit, with the possible application in a rectenna. The package inductance was reduced to 0.5 nH on each diode pin, as the devices are small and those values can be reached with advanced microwave circuit techniques, like flip chip mounting, striplines and printed passive components on the circuit board. The rectifier was tuned and simulated at a frequency of $f = 2.5\text{GHz}$. Power was $P_{in} = 18.4\text{W}$, $P_{out} = 16.4\text{W}$, its efficiency with ideal passive components is $\eta = 89\%$.

These results were compared to the performance of Rectennas, which were built to operate in the same frequency range. The biggest single rectenna found in literature operated at an output power of 6 W with an efficiency of 80 % at the same frequency [12]. Almost three times the power with one single rectifier could be achieved at very promising efficiencies in the simulation. Also the rectifier was chosen for its low component count and simplicity, making it even more attractive to be used in a rectenna array for the conversion of microwaves with a high power density to DC.

⁶Concretely measured was $c_f(V)$. This was done with a VNA whose measurement ports were elevated by an external DC voltage.

6. Measurement Setup

This chapter describes the measurement setup in the laboratory, used to benchmark and characterize the different rectifier prototypes. The RF to DC efficiency and the complex reflection coefficient are measured concretely. To get those values, at different power levels, the raw time domain data from a sampling oscilloscope is processed by a Matlab script.

6.1 Laboratory setup for pulsed power measurements

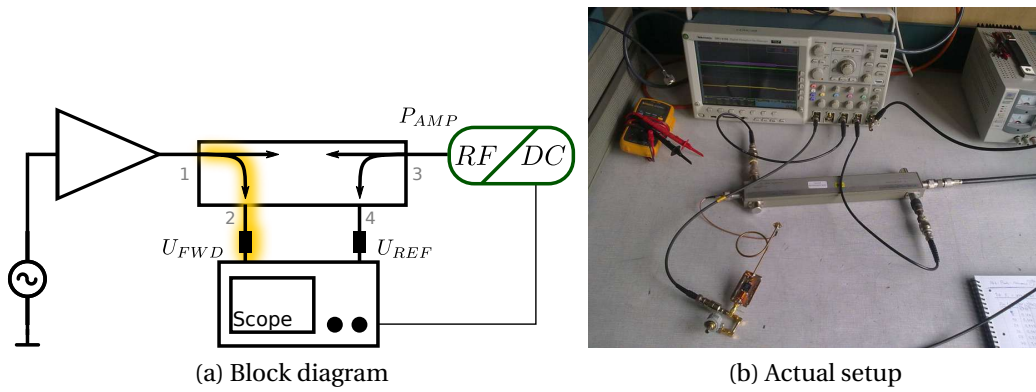


Figure 6.1: Laboratory setup for doing peak power measurements.

The setup used for pulsed power measurements can be seen in Figure 6.1a. It consists of a signal source, a small band pulsed RF power amplifier, a sampling oscilloscope, a broadband directional coupler and the actual device under test (DUT).

$P_{out} = 56dBm$	$P_{out} = 400W$	$G = 41dB$	$f_{out} = 202.5MHz$	$t_{on} = 1ms$	$t_{off} = 1s$
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Table 6.1: Technical data of the power amplifier.

The amplifier is a Cern proprietary design and originally used in linac 2, its technical data is summarized in Table 6.1. It is optimized for pulsed power and small bandwidth signals, its operating frequency is 202.5 MHz, which fits this application close enough. It allows output powers up to 56 dBm into a 50Ω load. The output power depends linearly on the input power with a nominal gain of 41 dB. The amplifier has been designed with robustness in mind. An internal circulator protects the output transistors from reflected power and allows the amplifier to operate with its outputs shorted or open.

The maximum on time of the device is 1 ms and the max. repetition rate 1 burst/s. A TTL input is used to trigger the RF burst. An internal control logic protects the amplifier from overloading. The trigger signal is generated by a Agilent mixed signal generator

Property [dB]	Calculation	$Fwd_{(1)}$	$Ref_{(1)}$	$Fwd_{(2)}$	$Ref_{(2)}$
Coupling	$20 \cdot \log(S_{21})$	20.98	20.97	19.53	19.52
Insertion loss	$20 \cdot \log(S_{31})$	0.15	0.15	0.24	0.24
Isolation	$20 \cdot \log(S_{41})$	52.40	56.86	53.09	54.57
Directivity	$20 \cdot \log(S_{41}/S_{21})$	31.42	35.92	33.55	35.04

Table 6.2: Characteristics of the directional coupler at $f_{(1)}$ and $f_{(2)}$

and distributed to the amplifier and the sampling scope. For all pulsed measurements a $1\mu s$ burst with a repetition rate of 1 burst/s is used.

The network analyzer (NA.) is used for practical reasons as the RF signal source with an adjustable power from $P_{GEN} = -80 \dots 17 dBm$. It is operating in continuous wave mode at 202.5 MHz, providing a single frequency signal as input drive for the amplifier. The NA is also used to determine the attenuation of the various components, which allows a calibrated measurement of the absolute input power level.

The sampling scope can acquire waveforms at up to 5 GS/s which allows to view signals with up to 500 MHz bandwidth. (Assuming at least 10 samples are needed to acquire one period). This not only allows measurement of the DC output voltage, which will be pulsed. Also it is fast enough to acquire the waveform of the actual RF voltage. By having the actual waveform data (and not just the envelope, like in usual time domain measurements) information about the phase and amplitude relation between forward and reflected wave can be extracted. Also the true RMS voltage can be calculated for nonsinusoidal waveforms. The scope has RF compatible inputs¹ with a characteristic impedance of 50Ω .

To measure incident and reflected power to the DUT, a broadband directional coupler is used. Its characteristics on the fundamental and harmonic frequency have been measured with the VNA and are shown in Table 6.2. As the device contains 2 separate 3 port couplers with internal termination resistors, all characteristics are measured for the forward and reverse direction. Using internal termination improves measurement accuracy, as the manufacturer already made best effort in minimizing reflected power from the internal isolated port. Table 6.2 shows a directivity of around 33 dB, this limits the dynamic range of the measurements. For example, if a forward signal of -33 dBm is observed on a frequency where also a reflected signal of 0 dBm is present, then the observed signal actually does not exist and is caused by the non ideality of the directional coupler.

6.2 Measurement methods

6.2.1 Efficiency

The efficiency is defined as the rectifiers DC output power P_{DC} divided by its available RF power P_{AMP} .

¹High impedance probe leads of prevalent scopes are a major source of trouble for RF measurements. Although the real part of the input impedance might be in the area of $1M\Omega$, the imaginary part is much lower and causes a intolerable level of capacitive loading on the circuit, severely influencing the measuring result or not allowing circuit operation at all.

A directional coupler connected to a sampling oscilloscope is used to measure U_{FWD} , then P_{AMP} is determined from this measurement. A scope is used as the on time of the amplifier is at most 1 ms , a mechanical or thermal RF power meter – measuring average power – would not respond fast enough for an accurate measurement.

For determining P_{AMP} it is critical to know the exact value of G_{ATT} which is the sum of:

- Coupling factor of the directional coupler.
- Cable attenuation between coupler and scope.
- Additional attenuator to protect the scope.
- Connection junctions.

If G_{ATT} is known, the RF power at Port 1 of the coupler – which is assumed to be the power going to the DUT – is related to it by Equation 6.1. It showed in measurements, that a deviance of 0.3 dB translates to a change of over 6% in the resulting efficiency value. In fact these 0.3 dB can easily be engendered by short laboratory cables. Thats why special care has been taken in measuring G_{ATT} of the **whole** setup with the NA. The critical path, forming G_{ATT} is highlighted in Figure 6.1a.

$$P_L = \frac{U_{RMS}^2}{50\Omega} \cdot 10^{G_{ATT}/10} \quad (6.1)$$

To determine the input power, U_{RMS} must be known, it is the root mean square value of U_{FWD} and calculated in Matlab by Equation 6.2.

$$U_{RMS} = \sqrt{\frac{1}{N} \sum_{n=1}^N U_{FWD}(n)^2} \quad (6.2)$$

The scope is set up to sample 100k points at a rate of 5 GS/s which gives a time window of $20\mu\text{s}$. The time domain data is exported to a PC and post processed in Matlab. A typical trace is shown in Figure 6.2. For the calculations, only a slice, from $6\mu\text{s}$ to $7\mu\text{s}$ is used, where the circuit is in steady state.

Obtaining the DC output power P_{DC} is straightforward as the output voltage is sampled by the scope. The load resistance is determined in disconnected state with a multimeter for every measurement and entered manually in the Matlab processing script. The output power and efficiency can be found from Equation 6.3. To get rid of quantization errors and residual AC contents, the output voltage is averaged over all samples in Matlab.

$$P_{DC} = \frac{U_{OUT}^2}{R_L} \quad \eta = \frac{P_{DC}}{P_{AMP}} \quad (6.3)$$

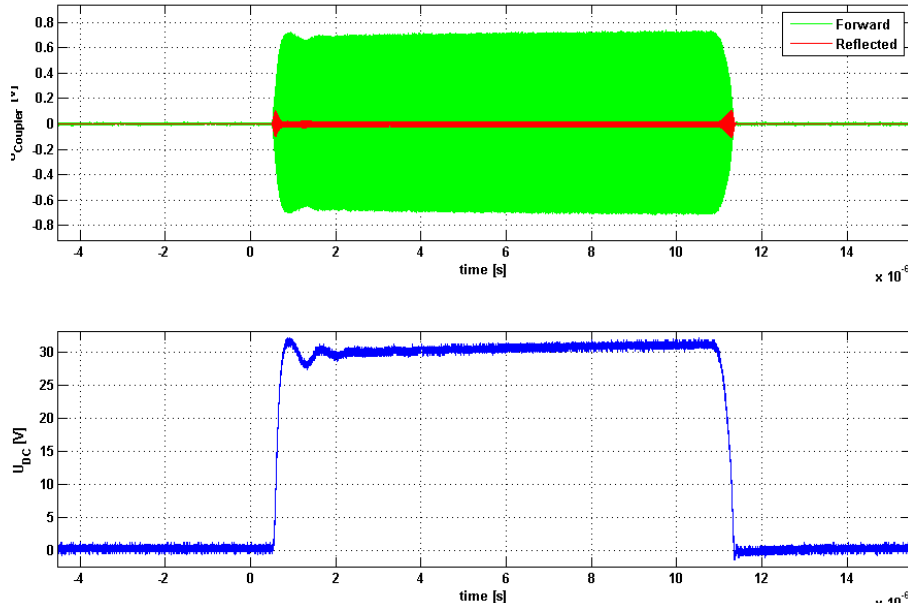


Figure 6.2: Typical waveform acquired by the scope.

6.2.2 Reflection coefficient

The complex reflection coefficient Γ is defined as the ratio of reflected and incident wave at the input of the DUT. As those waves are known, Γ can be determined from their relative phase and magnitude.

$$\Gamma = \frac{\hat{U}_{REF} \cdot e^{j\varphi_{REF}}}{\hat{U}_{FWD} \cdot e^{j\varphi_{FWD}}} \quad (6.4)$$

Visualizing Γ in a smith chart for different power levels provides an easy to interpret insight of the DUT behavior. This measurement setup actually corresponds to a vector network analyzer operating on a single frequency. It has the advantage over the VNA that it can measure with short pulses at very high powers, making it suitable for characterizing nonlinear devices.

Disadvantages – compared to the superhetrodyne concept of a regular VNA – are, that the oscilloscope needs to sample the input signal at its full RF bandwidth, this is nowadays only possible up to a few GHz with high end analog to digital converters. The broadband acquisition has a worse signal to noise ratio and less dynamic range compared to the narrow band measurements of a VNA.

Another disadvantage is the limited directivity of the used directional coupler, compared to a commercial VNA, where the internal directional couplers are of very high quality. This further limits the dynamic range of measurements.

Several steps are necessary for this measurement:

1. Forward and reflected wave of the DUT are acquired by the scope at different power levels. Each measurement consists of 100k samples per channel and is saved to a file on an USB stick.

2. The files are imported into Matlab, so that the waveforms U_{FWD} , U_{REF} and U_{DC} are available. Each file will give one point in the smith chart. The raw measured data can be seen in Figure 6.2
3. The transient response of the waveforms is clipped. The data from $3\mu s$ to $9.6\mu s$ (relative to the trigger pulse of the amplifier) is used, which gives around 33k points.
4. U_{FWD} and U_{REF} is processed by a fast Fourier transformation (FFT). It is done over $N = 2^{15}$ points. The sample rate is $f_s = 5GS/s$ giving a spectrum consisting of $N/2$ complex frequency bins and after Equation 6.5 a resolution of $\Delta f = 152.6kHz$.

$$\Delta f = \frac{f_s/2}{N/2} = f_s/N \quad (6.5)$$

To alleviate leakage effects in the spectrum, resulting from the finite number of samples, the time domain waveform is multiplied by a Chebyshev window. This means the resulting frequency domain data is convoluted with the windows spectrum. Spectral leakage is reduced substantially at the price of an reduced Δf . The resulting spectral bins have to be normalized to the average power of the window, as it introduces an energy loss due to tapering of amplitudes at the beginning and the end of the samples. The windows phase response is always linear², not changing the relative phase relationship of the results.

The final result of the FFT is shown in Figure 6.3a. Note that the noise floors are determined by quantization noise, the different levels result from different sensitivity settings of the oscilloscope³.

The forward voltage from the narrow band amplifier shows significant harmonics. They do not originate from the amplifier itself but are actually radiated harmonics from the rectifier. They travel towards the amplifier and get reflected on the internal narrow band circulator, which is high ohmic for harmonic frequencies. So they can be seen on the incident power measurement. For future measurements, this effect can be eliminated by putting a broad band circulator between amplifier and directional coupler, terminating the reflected power on harmonics from the rectifier.

5. In this case, only the (linear) reflection coefficient of the fundamental is calculated, to identify this frequency, a maximum search is done on the magnitude values of the U_{FWD} spectrum.

It is also possible to determine the reflection coefficient for any harmonic frequency. In this case the system would have to be excited at those frequencies and the particular frequency bin of interest needs to be chosen. Going this way

²A FFT window is naturally an even and real function in the time domain. This means it is symmetric to the y-axis and $x(n) = x(-n)$ holds true. Those kinds of time signals result in “zero phase” spectrums [41]. To avoid calculating with negative time indices, the window function is delayed so shifted to the right. Thus the resulting spectrum has a linear phase, its slope corresponding to the delay.

³In fact the forward voltage was measured with a scope setting of 500 mV/div, the reflected voltage with 50 mV/div. This means the later measurement is only affected by a tenth of the quantization noise level (assuming the scope samples with the same number of bits) resulting in an offset between the two noise floors of 20 dB.

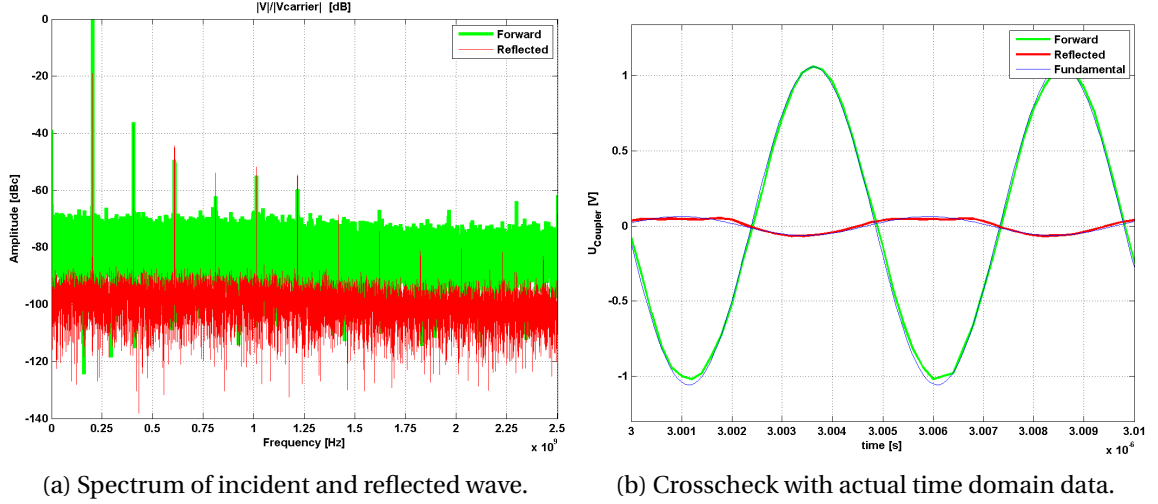


Figure 6.3: Example of the results obtained by the FFT operation.

further, a complete nonlinear model of the DUT can be created, more on this can be read in Chapter 6.2.3.

6. Magnitude and phase angle for U_{FWD} and U_{REF} of this particular bin of the spectrum is obtained. The FFT function in Matlab is defined as such that the absolute phase angle is relative to a pure cosine wave without phase shift. To check visually if the FFT method works and if its results are meaningful, the amplitude and phase values of the fundamental are taken and the sine wave is plotted over the actual measured waveform. This crosscheck can be seen in Figure 6.3b.

The raw value of $\Gamma_{M(1)}$ is calculated by Equation 6.4, dividing $U_{REF(1)}$ by $U_{FWD(1)}$.

7. The previously obtained calibration data and Γ_M is used to get the final Γ_{DUT} where systematic errors of the measurement setup are compensated. The method is explained in detail in Chapter 6.2.2.1.
8. Each point of Γ_{DUT} is drawn in a smith chart diagram.

Reflection coefficients for harmonic frequencies have not been considered in this work. Reasons are the mentioned reflections from the amplifiers output, the limited analog bandwidth of the scope (500 MHz) and the lack of calibration over a wide frequency range. Those problems can be solved without considerable effort in future work, which allows to convert the measurement data into a nonlinear x-parameter model, readily applicable in simulation software (see Chapter 6.2.3).

6.2.2.1 Error Model and Calibration

The calibration is based on a 3-term error model, described in [42]. The error model can be seen in Figure 6.4.

To get a mathematical model of the measurements error terms, we start looking at the values from the forward and reflected wave – a_1 and b_1 – that we see with the measurement setup. However, those are not the values of the actual DUT – a_{DUT} and b_{DUT} – we are interested in. They are altered by the error network in between. It suits as a model

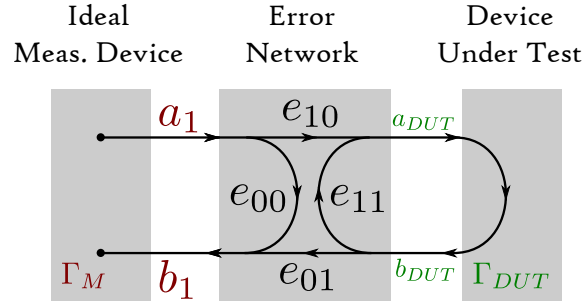


Figure 6.4: The adopted error model. The red values are measured, the green values are calculated.

for all systematic errors caused by nonidealities of the directional coupler, cables, and connecting junctions.

If the s-parameters of the error network are known, Γ_{DUT} can be calculated from the measured reflection coefficient Γ_M . Equation 6.6 shows the relations between the s-parameters of the error network and its in- and output waves

$$\begin{bmatrix} a_{DUT} \\ b_1 \end{bmatrix} = \begin{bmatrix} e_{11} & e_{10} \\ e_{01} & e_{00} \end{bmatrix} \cdot \begin{bmatrix} b_{DUT} \\ a_1 \end{bmatrix} \quad (6.6)$$

The relation between Γ_M and Γ_{DUT} can be written as

$$\Gamma_M = \frac{b_1}{a_1} = e_{00} + \frac{e_{01}e_{10}\Gamma_{DUT}}{1 - e_{11}\Gamma_{DUT}} \quad (6.7)$$

The error network is assumed to be reciprocal, this means the gain in forward and reverse direction is assumed to be equal ($e_{10} = e_{01}$), simplifying the equation. Now the s-Parameters of the error network can be determined by connecting 3 devices with known reflection factors to the setup.

Γ_O , Γ_S and $\Gamma_M = 0$ are the known reflection coefficients of a commercial network analyzer calibration kit⁴. Only the match is assumed to be ideal, simplifying this calculation. Measuring each of these standards with the setup we get M_O , M_S and M_M . By replacing Γ_M and Γ_{DUT} in Equation 6.7 we get a system of 3 equation that can be solved for the 3 error parameters:

$$e_{00} = M_M \quad (6.8)$$

$$e_{10} = \frac{(\Gamma_O - \Gamma_S)(M_O - M_M)(M_S - M_M)}{\Gamma_O\Gamma_S(M_O - M_S)} \quad (6.9)$$

$$e_{11} = \frac{\Gamma_S(M_O - M_M) - \Gamma_O(M_S - M_M)}{\Gamma_O\Gamma_S(M_O - M_S)} \quad (6.10)$$

⁴It shall be noted that those devices don't reflect a perfect open or short circuit. Their value (and thus price) lies rather in the fact that their electrical properties are very stable, don't change over time and have been accurately measured by the manufacturer for each device. They are delivered with a printout or a floppy containing the measurements results.

Having determined the 3 s-parameters by the calibration procedure, those can be used to compensate the systematic error from any measurement at one frequency. Solving Equation 6.7 for Γ_{DUT} gives Equation 6.11, a formula for the corrected reflection coefficient.

$$\Gamma_{DUT} = \frac{M_{DUT} - e_{00}}{e_{10} + e_{11}(M_{DUT} - e_{00})} \quad (6.11)$$

It shall be noted that this calibration procedure is only valid for one particular frequency, meaning that it has to be processed for the fundamental and all the harmonic frequencies that are considered in the measurement separately. The actual measurement of the calibration standards has to be done only once, as the FFT function extracts the considered frequency bins from the time domain response. This rises an important point, to calibrate the system on harmonic frequencies, it must be excited on those frequencies too. Thus a broadband input signal with enough energy on the frequencies of interest should be used for the calibration procedure. Good examples test signals are step or impulse functions, containing energy on a very broad frequency spectrum.

6.2.2.2 Verification and accuracy

To verify that the measurement setup gives meaningful results, several linear test devices have been measured with a conventional VNA and the scope setup. Attenuators, connected to various length of cables with their ends open or shorted have been used as test devices to obtain various reflection coefficients, they are shown in Figure 6.5a. The results are compared in one smith chart shown in Figure 6.5b.

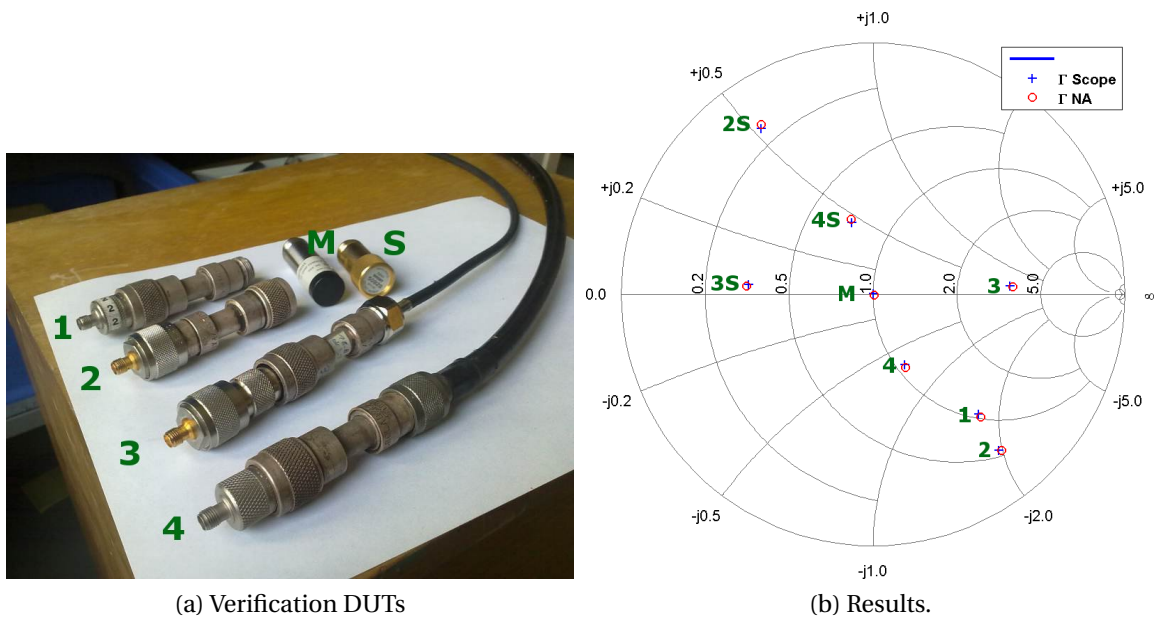


Figure 6.5: Comparison of the scope measurement setup with a commercial VNA.

6.2.3 X-Parameters

The reflected waves – measured in Chapter 6.2.2 at the fundamental and harmonic frequencies, at different power levels – can be used to create a one port X-parameters⁵ model of the DUT.

X-parameters are a superset of classical S-parameters and provide the necessary mathematical framework to measure, model and simulate nonlinear systems [43]. Models can show particular characteristics of nonlinear systems, like the generation of harmonics or intermodulation distortion, which linearized S-parameters are not able to. For small input signals they converge to classical small signal S-parameters.

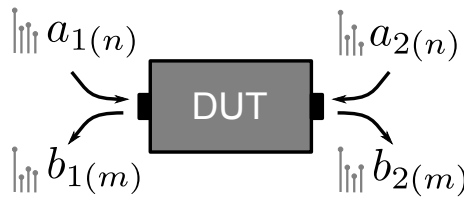


Figure 6.6: Power waves on a two port.

As in S-parameters, power waves are described going in (a) or out (b) from ports of the DUT. The general definition for a wave going in or out port 1 is

$$a_1 = \frac{V_1 + I_1 Z_0}{2\sqrt{Z_0}} \quad b_1 = \frac{V_1 - I_1 Z_0}{2\sqrt{Z_0}} \quad (6.12)$$

and has the unit of $\sqrt{\text{Power}}$. More practically, if a port is internally terminated by a resistor with the systems characteristic impedance Z_0 , there will only be an incident wave (a_1) and the voltage measured across this resistor U_R relates to the incident power wave by $a_1 = U_R / \sqrt{Z_0}$.

S-Parameters are now simply defined as ratios of one incident and one reflected wave. Which waves are used is indicated by an subscript, for example the transmission characteristic is defined as $S_{21} = b_2 / a_1$. Normalizing by the incident wave makes the S-Parameter invariant of the probing signal or measurement setup and only dependent on the DUT.

This way all possible combinations of waves at the DUTs ports are measured one after the other⁶. Using linear superposition allows now to calculate the response of the system on any port (b_y) to a sinusoidal excitation on any (or even multiple) other ports (a_x).

The problem is that s-Parameters are nothing more than complex gain factors. While a wave passes the DUT from one to the other port, its magnitude or phase might change, but it still needs to be a single sinusoidal wave at the output. S-parameters provide a linear small signal model of the DUT, that can't reproduce real life nonlinear effects like distortion or intermodulation, creating harmonics frequencies. To get around this problem, X-parameters have been invented.

⁵X-parameters is a registered trademark of Agilent Technologies.

⁶It is important to terminate all remaining ports by the systems characteristic impedance to avoid any reflected waves from ports.

While S-parameters are a **linear** describing function in the sense of

$$b_2 = f(a_1, a_2, \dots) \quad (6.13)$$

X-parameters expand this to consider harmonics. A second subscript is introduced, defining the harmonic index in brackets.

$$b_{2(1)} = f(a_{1(1)}, a_{1(2)}, \dots, a_{2(1)}, a_{2(2)}, \dots) \quad (6.14)$$

The involved math simplifies, as the system is assumed to be time invariant, any delay on the input signal will not change its characteristics. Also only one of the incident signals is assumed to be a large signal, able to cause distortion. This is the reference signal $A_{1(1)}$ which serves as a phase reference for all other measurements. For any other input signals the principle of superposition is assumed to hold, as they have to be small enough (they are called tickle signals in the actual measurement). Also all X-parameters are dependent on input amplitude of the reference signal.

So in the end, the function from Equation 6.13 – giving the definition of S-parameters – is realized by:

$$b_p = \sum_q S_{pq} a_q \quad (6.15)$$

The function from Equation 6.14 – giving the definition of X-parameters – is realized by:

$$b_{p(m)} = \sum_{qn} S_{pq,(m)(n)}(|a_{1(1)}|) P^{(m)-(n)} a_{q(n)} + \sum_{qn} T_{pq,(m)(n)}(|a_{1(1)}|) P^{(m)+(n)} a_{q(n)}^* \quad (6.16)$$

The subscript p is the output port, m the output harmonic index. Both sums go over each input port q and each input harmonic n.

The first sum contains a classic small signal S-parameter for the link between output and input ports. It is a function of the input power on port $a_{1(1)}$. P is the Phasor and defined as

$$P = e^{j\Phi(a_{1(1)})} \quad (6.17)$$

It is needed to make the model time invariant and allows $\Phi(a_{1(1)})$ to be a phase reference for all harmonics on all ports. Also it is defined in such a way that the large signal component does not have an imaginary part, simplifying calculations.

The second sum contains the T term, which actually is a function, depending on amplitude and phase difference of the input to the reference signal. It is introduced, as the modeled system might change its behavior with the phase relation of $a_{q(n)}$ to the reference $a_{1(1)}$. This is the first difference to S-parameters, where the absolute phase of one input signal is not significant.

As we only do measurements on one single port and are only equipped with a single tone source, the generalized theory shall not be longer exhausted. More on this can be

found in [43]. For this special case, the generalized theory simplifies a lot, the T term gets zero and the equation is:

$$b_{1(m)} = S_{11,(m)(1)} (|a_{1(1)}|) P^{(m)-1} a_{1(1)} \quad (6.18)$$

Or to measure one actual X-Parameter:

$$S_{11,(m)(1)} (|a_{1(1)}|) = \frac{b_{1(m)}}{a_{1(1)}} e^{-j\Phi(a_{1(1)})[(m)-1]} \quad (6.19)$$

That means we get a series of parameters for different harmonic indexes m, each one has to be measured as a function of input amplitude $|a_{1(1)}|$. The factor $b_{1(m)}/a_{1(1)}$ relates to the reflection coefficient Γ for the fundamental and harmonics calculated in Chapter 6.2.2 by the Matlab script. The exponential term normalizes the phase of the harmonics to that of the fundamental. All phases are relative to $\Phi(a_{1(1)})$.

This shows, that it is easy to create a one port X-Parameter model by the measured data from the setup in Chapter 6.2.2. The model would describe the complete nonlinear behaviour of the rectifier on its input port. One very interesting use for this model is to import it into the simulation software Agilent ADS – which supports X-Parameters innately. This would allow the simulation of a whole array of rectifiers, together with the characteristics of the RF power divider, circulators and other microwave components. There is a potential to simulate the whole energy recovery system and analyze its impact on the travelling wave cavities in detail.

7. Prototypes and measurement results

Several prototypes of rectifiers have been built in the laboratory. They serve the purpose of relating the theoretical results obtained mathematically and by simulation to measurements in real life. For this reason one non resonant and two resonant rectifiers have been designed, built, operated and measured in the laboratory.

Each prototype has been tested for at least 30 minutes with pulse lengths of 1 ms at the maximum power. This was done to ensure that the rectifiers are operating reliably. No performance degradation was observed during those tests, making continuous wave operation only a matter of heat removal.

7.1 Prototype 0) Simple full wave rectifier

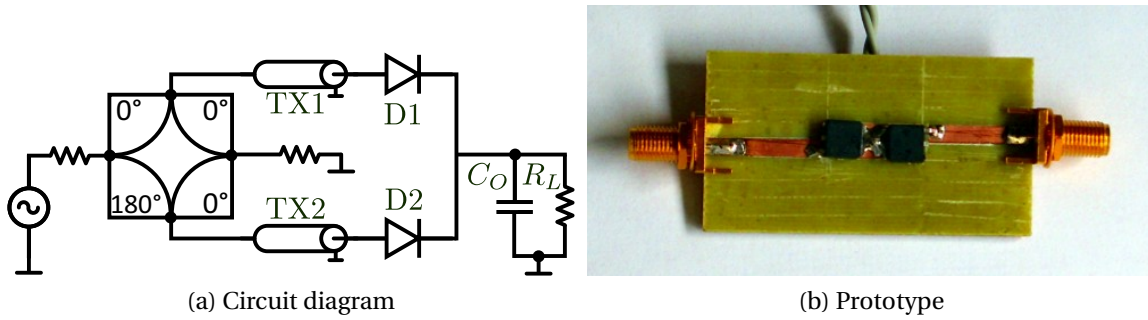


Figure 7.1: Simple full wave rectifier.

This rectifier is the first prototype which has been built to get a practical grip on the engineering challenges with nonlinear, real life diodes, including their parasitics. It also helped in understanding nonlinear RF circuits and to develop a suitable measurement setup.

It consists of 2 diodes that are connected to 2 RF signals with a phase shift of 180 degree. This corresponds to the classical center tapped transformer coupled full wave rectifier from Chapter 4.5.3. In this case the phase shift has been obtained by a commercial 180 degree microwave hybrid.

The diodes are STPSC406B silicon carbide schottky diodes, their maximum ratings are $V_{rev} = 600V$, $I_{fwd} = 4A$. Their junction capacitance is specified as $c_{J1} = 140pF$ at 1V.

7.1.1 Construction

No measures for impedance matching or resonant operation have been taken. D1 and D2 are soldered on a printed circuit board (pcb) and connected to SMA connectors by controlled impedance microstrip lines. Those are traces with a specified geometry

over a groundplane and behave like transmission lines. To design them it is essential to know the effective phase velocity of the line.

The phase velocity for a electromagnetic wave equals the speed of light (c) in vacuum. If the wave travels in a medium with the dielectric constant ϵ , the phase velocity becomes:

$$v_c = \frac{c}{\sqrt{\epsilon}} \quad (7.1)$$

If the phase velocity, or respectively ϵ is known, a relation between electrical and physical length of a transmission line can easily be established:

$$l_{\text{phy}} = \frac{l_{\text{el}}}{\sqrt{\epsilon}} \quad (7.2)$$

The problem is, in a microstrip line the waves travel partly in the dielectric below and partly in the air above the circuit trace. Also fringing¹ effects have to be considered, which leads to complicated empirical formulas to determine an effective ϵ_{eff} which describes uniformly the entire space.

Equation 7.3 is one of the more accurate ones, it has been taken from [44] and used to design the microstrip lines. In the equation, $u = w/h$ is the ratio of trace width to the thickness of the circuit board.

$$\epsilon_{\text{eff}} = \frac{\epsilon + 1}{2} + \frac{\epsilon - 1}{2} \cdot \left(1 + \frac{10}{u}\right)^{-ab} \quad (7.3)$$

$$a = 1 + \frac{1}{49} \ln \left[\frac{u^4 + (u/52)^2}{u^4 + 0.432} \right] + \frac{1}{18.7} \ln \left[1 + \frac{u^3}{18.1} \right] \quad (7.4)$$

$$b = 0.564 \left(\frac{\epsilon - 0.9}{\epsilon + 3} \right)^{0.053} \quad (7.5)$$

With the ϵ_{eff} , the electrical length of the transmission line can be calculated. The formula for its characteristic impedance, dependent on w , h and ϵ_{eff} is given in Equation 7.6.

$$Z_0 = \frac{376.82}{2\pi\sqrt{\epsilon_{\text{eff}}}} \ln \left[\frac{f(u)}{u} + \sqrt{1 + \frac{4}{u^2}} \right] \quad (7.6)$$

$$f(u) = 6 + (2\pi - 6) \exp \left[- \left(\frac{30.666}{u} \right)^{0.7528} \right] \quad (7.7)$$

The formulas have been transferred to an Excel spreadsheet which allowed to determine the trace width $w = 3.34\text{mm}$ to get a $Z_0 = 50.01\Omega$ impedance for a circuit board thickness of $h = 1.5\text{mm}$ and a relative dielectric constant of $\epsilon = 3.5$.

The traces have been cut out by hand from the circuit board.

The diodes are terminated by C_O which consists of 2 x 470 pF ceramic chip capacitors. The DC output can be loaded with a variable resistor R_L with a range of 0 – 100 Ω .

¹Fringing fields can be observed when looking at the electric field of a parallel plate capacitor. The field between the plates is uniform, the field lines are orthogonal to the plates. At the flanges of the plates, the field lines extend out into open space and form a arcs. These extending field lines are hard to calculate and called fringing fields.

$P_{AMP}[W]$	$R_L[\Omega]$	$U_L[V]$	$P_L[W]$	$\eta[\%]$
50	38.7	14.0	5.0	10.0
100	19.9	22.6	25.7	25.7
200	19.5	35.5	63.3	31.7

Table 7.1: Measured efficiency of prototype 0

7.1.2 Simulation

The diode model has been taken from the manufacturers website. Inductor L14 - L17 model the parasitic elements of the diodes, including their package inductance. The 2 out of phase outputs of the microwave hybrid have been modelled by two voltage sources, each with half the amplitude. Total input voltage is 283 V resulting in an available power of 200 W. As the DC current flows through the microwave hybrid², its DC resistance has been measured and included as R10 in the simulation. The simula-

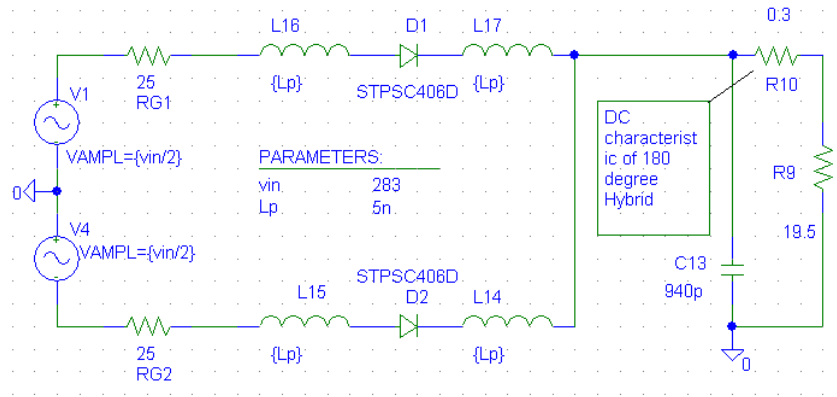


Figure 7.2: PSpice simulation of the prototype.

tion resulted in an DC output voltage of $U_{R9} = 33V$, giving 56 W of DC power. This corresponds to an efficiency of 28 %.

7.1.3 Measurements

The results of the efficiency measurements of the real device are shown in Table 7.1. The load resistance has been varied until a minimum of reflected power has been reached. As there is no impedance matching whatsoever, even after optimization the reflected power was more than 60% of the incident power.

Harmonic radiation is mostly filtered and dissipated in the small band hybrid and its termination resistance.

7.1.4 Conclusion

The measurements showed that impedance matching to the diodes is critical for efficient RF operation. Especially as the diode shows a large c_j , which lowers its input impedance far beyond 50Ω. Nonetheless the experiment showed that switching diodes –

²It should be noted that not all hybrids are designed to conduct DC current. Some are not conductive at all. Especially those based on ferrite conduct DC but the magnetic materials can saturate which leads to distortion and other effects on the RF signals.

not designed for RF operation – still can be used for an 200 MHz rectifier. The simulation results agree with the measurement, showing that a transient simulation in PSpice can be used to predict a nonlinear RF circuits behavior.

7.2 Prototype 1) Current output series resonant rectifier

Prototype 1 has been built after the simulation results from Chapter 4.5 suggested it would be the most suitable design. Also improvements from the first prototype are incorporated.

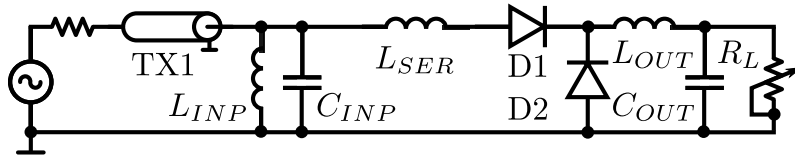


Figure 7.3: Prototype1: Schematic

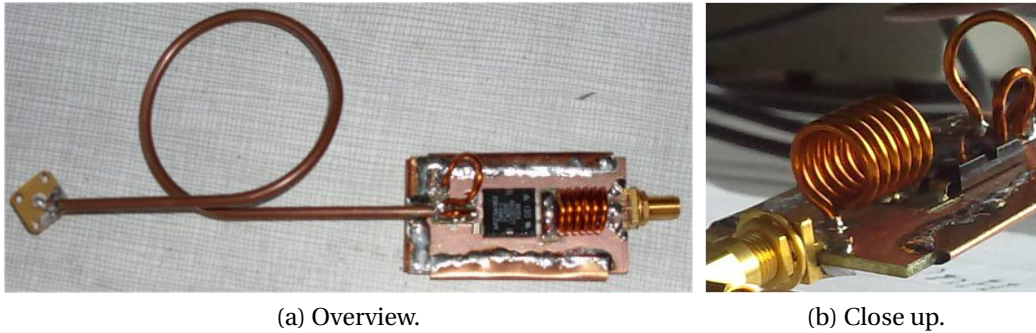


Figure 7.4: Prototype1: Series resonant current output rectifier.

7.2.1 Construction

It uses a GS150TA25110 GaAs Schottky Diode from IXYS. The package contains 3 diodes with a common collector connection. One of them stays unused. For impedance matching, a simple quarter wave line is used. The output load R_L is an adjustable resistor (potentiometer). Wire wound inductors, SMD capacitors and a copper clad board – with circuit traces cut by hand – have been used for the construction.

The simulation of the circuit suggested an input impedance of $Z_{IN} = 16.2\Omega$ at 200 W input power (See Chapter 4.4). To get optimum power transfer, this has to be matched to the 50Ω output of the power amplifier. A rigid transmission line, with an electrical length of a quarter wavelength and $Z_0 = 25\Omega$ is used. It transforms the rectifiers impedance to $Z_0^2/Z_{IN} = 38.5\Omega$ at 200W. This is a compromise which has been accepted as no transmission line with a more suitable Z_0 was available in the laboratory. Looking at Figure 4.19 it should give the best match at around 40 W input power. The line was cut to an electrical length of $\lambda/4 = 37.5\text{cm}$ by connecting it to a calibrated VNA, shortening the other end and cutting small pieces off. $\lambda/4$ is reached as the VNA shows an open circuit at 200 MHz in the Smith chart.

The inductivities have first been estimated by an laboratory formula, then wound with magnet wire and soldered to an SMA receptacle. This allowed measurements with the calibrated VNA and showed the influence of parasitic capacitance very clearly. This way different inductor shapes have been tested. In the end it was possible to build the output inductor $L_{out} = 143\text{ nH}$ – seen in Figure 7.4b – with a self resonant frequency of 578 MHz. Above this frequency the inductor starts behaving like a capacitor. The advantages of wire wound air inductors is, that they can be tuned by hand very easily. By bending the wires and expanding or contracting the coils, their inductance can be adjusted.

7.2.2 Measurements

The efficiency and reflection coefficient has been measured for a series of different input power levels, like described in Chapter 6.2. For the first measurement, R_L was left constant, set to an value that gives least reflection at around 200W. For the second measurement R_L was adjusted for least reflection at each power setting.

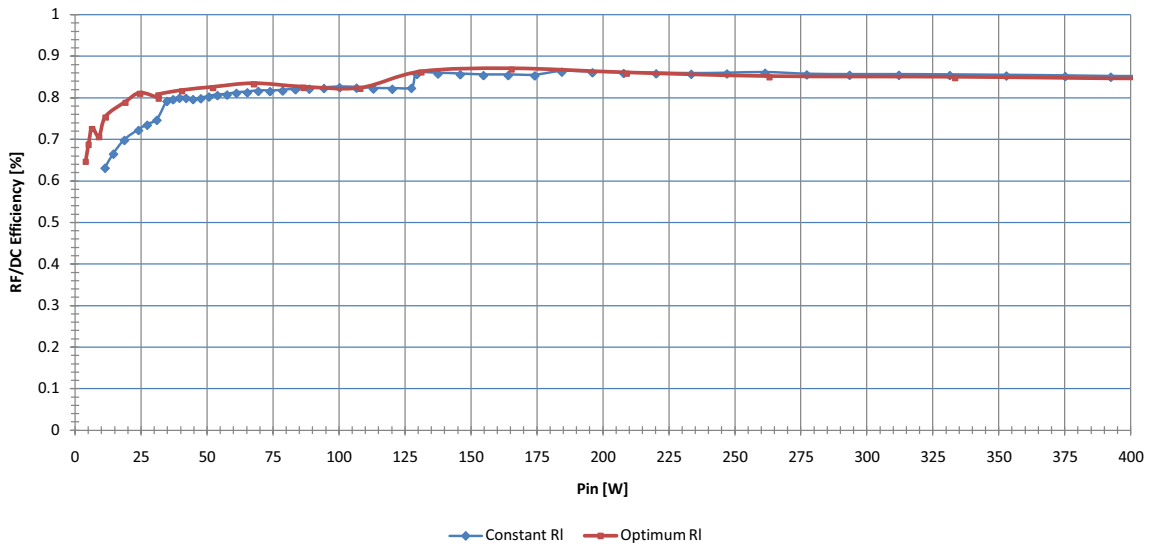
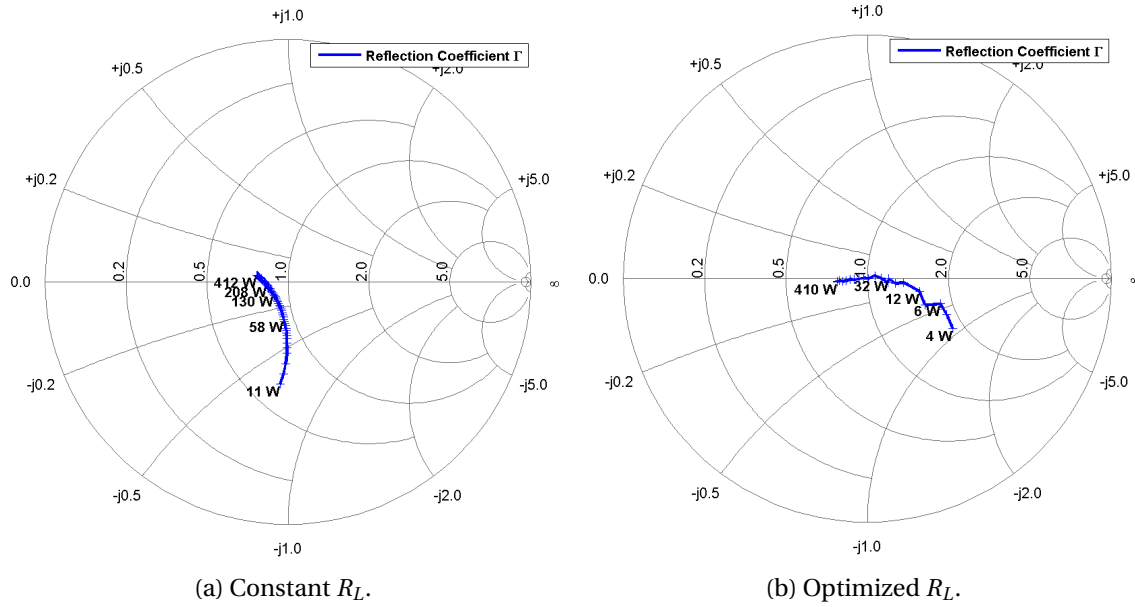


Figure 7.5: Prototype1: Energy conversion efficiency over input power.

The achieved efficiency curves can be seen in Figure 7.5. For both measurements the curve is reasonably flat. Even with a constant R_L , the rectifiers efficiency stays above 80 % for an input power range from 50 to 400 W. Peak efficiency is reached at 165 W with 86.8 %.

Figure 7.6a shows the rectifiers complex reflection coefficient Γ for constant R_L . The effect of the diodes voltage dependent c_J can be seen clearly. For small input powers the voltage across the shunt diode D2 is small, c_J is big and the input behaves capacitive. As the power increases c_J gets smaller and the input impedance behaves less capacitive, forming a trajectory that resembles the constant conductance circles of the smith chart and goes upwards.

Figure 7.6b shows Γ for the case where R_L was optimized for each measurement. It can be seen clearly, how the rectifiers input reactance can be brought to zero by changing R_L . The voltage across the shunt diode D2 is set just big enough that its c_J is compensated by the inductances in the circuits RF loop. It can also be seen that the reflected power is zero at around 40 W. This confirms that the impedance match from Chapter

Figure 7.6: Prototype1: Reflection coefficient Γ over input power.

7.2.1 works as expected and the rectifiers input impedance was predicted correctly by the simulation.

The spectrum of incident and reflected voltage can be seen in Figure 7.7. All values are normalized to the input amplitude of the fundamental. Even with the simple LC input filter, all reflected power – on the fundamental and on every harmonic – is below -20 dB, which is acceptable for the application in the SPS. Note that the forward power also shows harmonics, they are caused by the small band circulator – built into the amplifier – reflecting harmonics back towards the DUT.

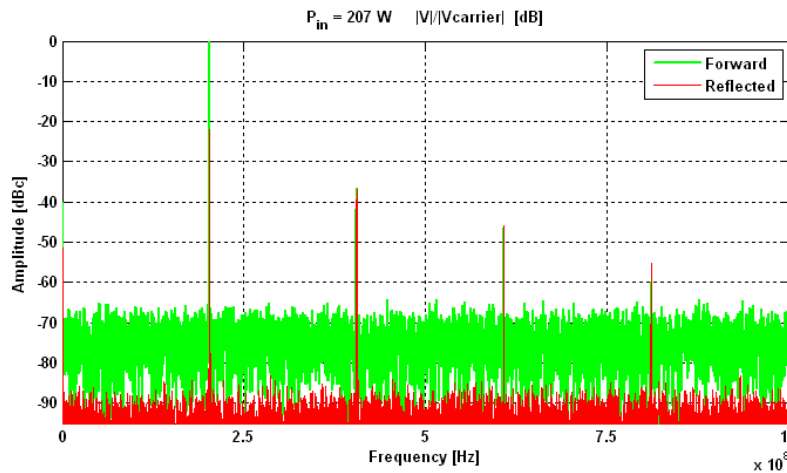


Figure 7.7: Spectrum of incident and reflected voltage for prototype 1.

7.2.3 Improved impedance match

Attempts were made to improve the impedance matching of the rectifier at higher powers. At 200W $Z_{in} = 42.1\Omega - 0.7j\Omega$ was measured. So the bare rectifier – before the $\lambda/4$ line

– shows an input impedance of $Z_0^2/Z_{IN} = 14.8\Omega$. To match this, an impedance transformer with 2 series transmission lines – having different Z_0 values – was designed like described in Chapter 4.4 and used to replace the former $\lambda/4$ line.

The first line is defined as: $Z_{01} = 25\Omega$, $l_{el} = 0.1623\lambda = 243.5mm$ and has been trimmed to length with the VNA. 50 Ω SMA connectors are soldered to the end which connects to the second line. The second line is defined as: $Z_{02} = 50\Omega$, $l_{el} = 0.0364\lambda = 54.6mm$ and it showed that the 2 50 Ω SMA connectors – connecting the first line to the circuit board – have already an electrical length in that area. So no explicit second line was used.

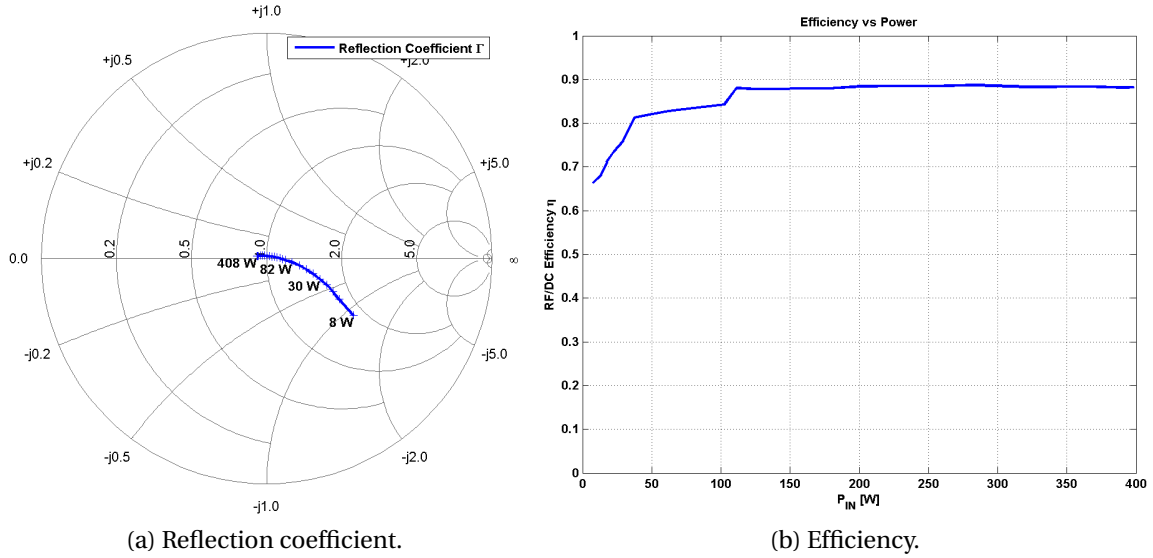


Figure 7.8: Prototype 1: Improved impedance match.

The results can be seen in Figure 7.8. The impedance is matched very well at higher powers. Also the trajectory in the smith chart is less steep, resulting in a good match over a wide input power bandwidth. This is a direct result of the second order matching circuit.

7.2.4 Conclusion

This prototype showed successfully, that high power RF can be converted to DC very efficiently with currently available diodes. The special low inductance package of the GS150TA25110 diode allows to tune the rectifier into resonance by a carefully adjusted L_{ser} . The simulated and real input impedance shows a good concurrence. Advanced impedance matching allows a very good power transfer at a broad range of input powers. The load resistance does not have to be adjustable, as the adjustment bears almost no efficiency advantage for power levels > 50 W. Best match is obtained at 225 W, maximum efficiency at 284 W with 88.7%.

It shall also be noted that the material cost for this prototype were < 20 €, making this architecture very attractive for deploying it in large rectifier arrays.

7.3 Prototype 2) Improved construction

With the results from Prototype 1, a final rectifier has been designed, using a commercially made 2 layer printed circuit board (pcb) and the layout program Eagle. This

rectifier has been designed for continuous wave (CW) operation, using high quality low loss components and taking care of heat dissipation.

7.3.1 Construction

The design of the impedance match has been kept the same as in Chapter 7.2.3, although now TX2 has been realized by the SMA connectors and a short trace of 50Ω microstripline (following the same design procedure as in Chapter 7.1).

The inductors were designed as spiral traces on the board, those inductors are very easy to manufacture, essentially cost nothing and are more repeatable than wire wound coils. Also they show less losses because of a large surface area³. The empirical design formulas from [45] have been used to determine the spirals geometry. No groundplane was used as the induced RF current would introduce losses and lower the Q of the inductor⁴. Also this allows to put metal shielding near the inductor afterwards to tune its value. The spirals were drawn in Eagle, special care has been taken to minimize coupling between the inductors and ensure there is no overlap (top and bottom side) between them.

MCM capacitors from Cornell Dubilier, made out of mica material – especially suited to conduct large RF currents and having minimum losses – are used for the input and output filters. They can stand the current stress of CW operation.

The PCB has been etched, drilled and electroplated by a professional PCB design house.

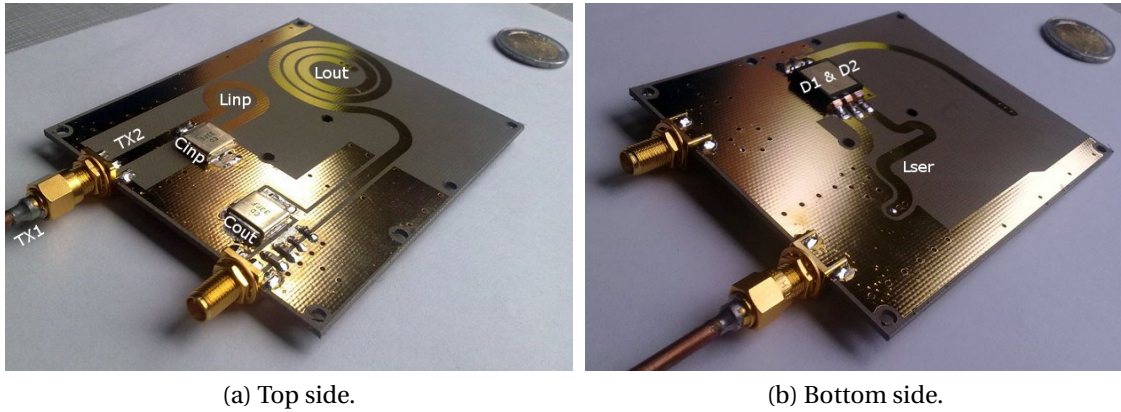


Figure 7.9: Prototype2: Printed circuit board design.

7.3.2 Measurements

The characteristic impedance and electrical lengths of TX1 and TX2 were checked in the finished circuit by time domain reflectometry (TDR) with a VNA. The instrument

³At RF frequencies the skin effect always has to be considered in conductors. In fact at 200 MHz, most of the current flows only in the top 5μm of the copper trace. The PCB has a copper thickness of 35μm which is more than enough to support the current flow, increasing the thickness will not increase trace resistance. On the other hand, the traces have very large surface area compared to the magnet wires used to form the coils of prototype 1, this promises less resistive losses and also better heat dissipation suiting the CW operation well.

⁴A printed inductor L_1 with ground plane can actually be seen as transformer with shorted secondary winding. The current flowing in the groundplane creates a mutual inductance $L_2 = L_1$, coupling negatively to L_1 . The result is that the inductance is reduced to $L'_1 = L_1(1 - k)$ where k is the coupling factor.

runs a frequency sweep, obtaining the reflection coefficient of the circuit in the frequency domain. Then it uses inverse Fast Fourier Transformation (iFFT) to calculate the virtual step response over time of the input. Every change in impedance will reflect some power and can be seen as a change of reflection coefficient at a specific time. The electrical length of the transmission line can be calculated by measuring the time difference between two changes in reflection coefficient. This is the round trip time of the line and related to its electrical length by:

$$l_{el} = \frac{t_{rtt} \cdot c}{2} \quad (7.8)$$

Z_0 can be determined from the actual value of the reflection coefficient on the y-axis. Impedance is related to reflection coefficient in a 50Ω system by the well known formula

$$Z = 50\Omega \frac{1 + \Gamma}{1 - \Gamma} \quad (7.9)$$

The advantage is, that multiple components connected in series can be easily distinguished by the corresponding offset on the time axis. Also the circuit is measured including all connection junctions, that might add additional electrical length, disturbing the impedance match.

The TDR measurement can be seen in Figure 7.10a. After the VNA has been calibrated as usual, a reference measurement, with an open test line, has been done. This produced the top trace, first showing Γ to be zero, then rising to one at $t=0$ because of the reflections from the open line end. To measure any time differences, the point in the middle of the slope is used as reference. The lower trace represents the actual measurement of the rectifiers input.

First a very short piece of 50Ω line can be seen, originating from the SMA female to female adapter used to connect the circuit. Then the reflection coefficient drops to $\Gamma = -0.333$ for $t_{rtt} = 1.598ns$, indicating a piece of line with $Z_0 = 25.0\Omega$ and $l_{el} = 239.5mm$. This corresponds to TX1, designed for $Z_0 = 25\Omega$, $l_{el} = 243.5mm$.

Then Γ rises again to zero, indicating a short piece of 50Ω line. It is $l_{el} = 64.5mm$ long. This represents the microstrip line TX2 and the 2 SMA connectors connecting it to TX1. The designed value for perfect match was $l_{el} = 54.6mm$ so the final value turned out a bit to long.

At the end Γ slightly rises over time. This is caused by the capacitor of the input filter, it looks like a short circuit in the beginning and then starts charging and slightly increasing its voltage.

Next the 2 inductors L_{ser} and L_{inp} were measured. A SMA receptacle was first measured with a calibrated VNA. Its electrical length was internally compensated, so the VNA shows only an open circuit. The receptacle was soldered to the circuit board, directly connecting to one of the inductors. The VNA now measures the reflection coefficient over frequency. It also shows impedance, calculated internally by Equation 7.9. A marker is used to read out the reactance at 200 MHz, seen in Figure 7.10b. The results are summarized in Table 7.2.

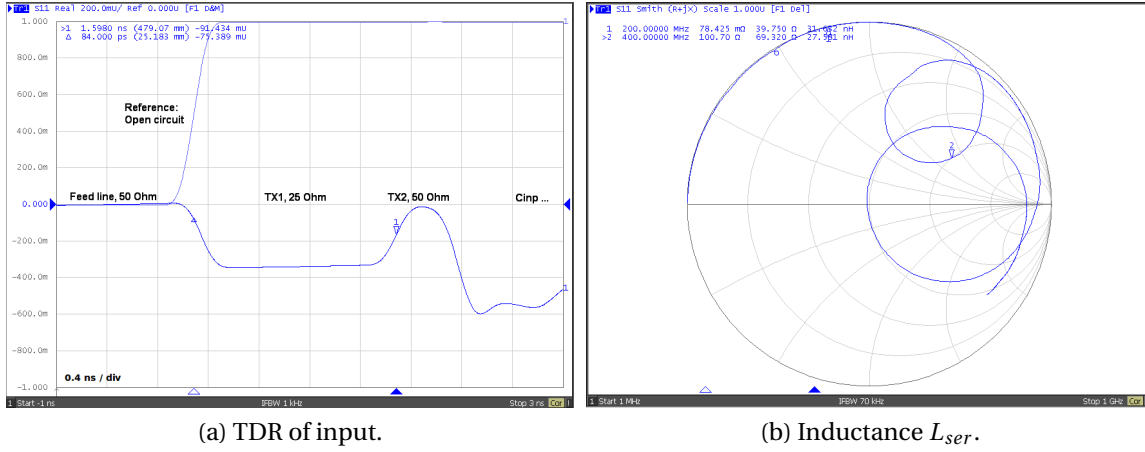


Figure 7.10: Prototype2: Measuring components.

	target	measured	corrected
L_{inp}	19.2 nH	24.8 nH	-
L_{ser}	14.0 nH	31.6 nH	21 nH

Table 7.2: Results of measuring the inductors.

Unfortunately did both inductors turn out bigger than expected. For L_{inp} this is not a problem as the shift in resonance frequency is small compared to the filters transmission bandwidth.

But the abundant L_{ser} puts the circuit into a high impedance mode of operation. A smaller c_J is needed to still meet the resonant condition, resulting in a higher output voltage for the same input power. Assuming R_L is adjusted to obtain resonance, it has to be set to a high ohmic value (see also Chapter 4.3).

Correction attempts for L_{ser} have been done by soldering copper sheet over the trace, increasing its width and also shorting the meander shaped part, which both reduced inductance to an acceptable but still not optimum value.

The results of the pulsed power measurement can be seen in Figure 7.11. A good impedance match could only be achieved at high powers, this has been already expected and is a direct result of the oversized L_{ser} . Only with a high voltage across the diodes, their c_J is small enough for resonance with L_{ser} at the operating frequency. Minimum reflection is achieved at 364W input power.

On the other hand, still very good efficiency values could be obtained. The maximum is at 224 W input power with 88.0%. The efficiency is lower than in Prototype 1, it is spoiled by bad input match at lower powers and naturally reduced because of conduction losses in the diode at higher powers.

The spectrum of the radiated harmonics can be seen in Figure 7.12. As the impedance match is not optimal, the reflected power on the fundamental exceeds -20 dB. The harmonics are well suppressed below -40 dB of the incident power.

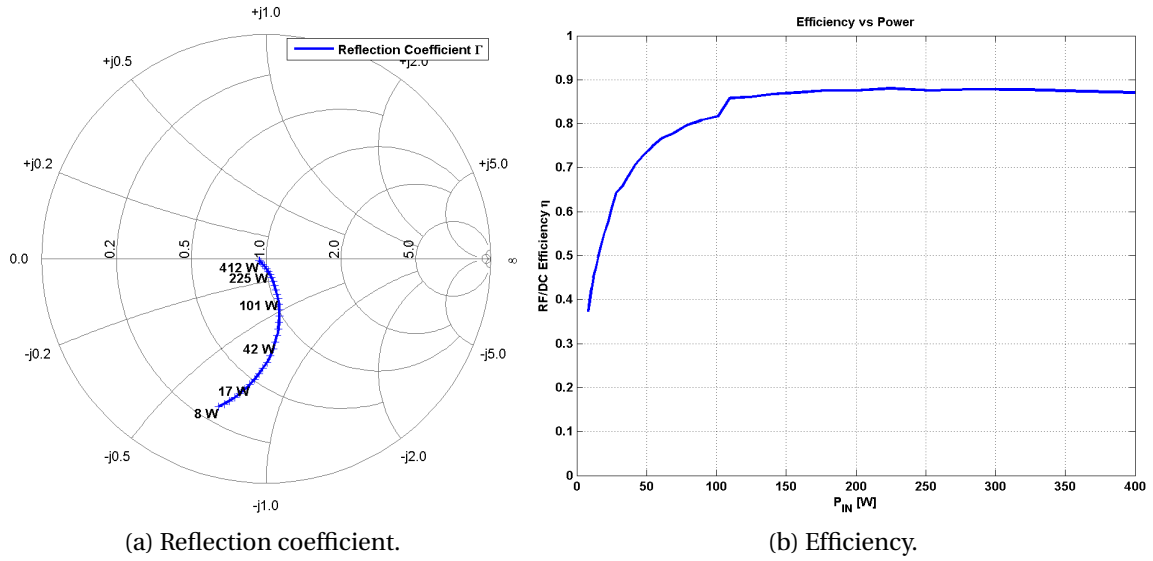


Figure 7.11: Prototype2: Printed circuit board design.

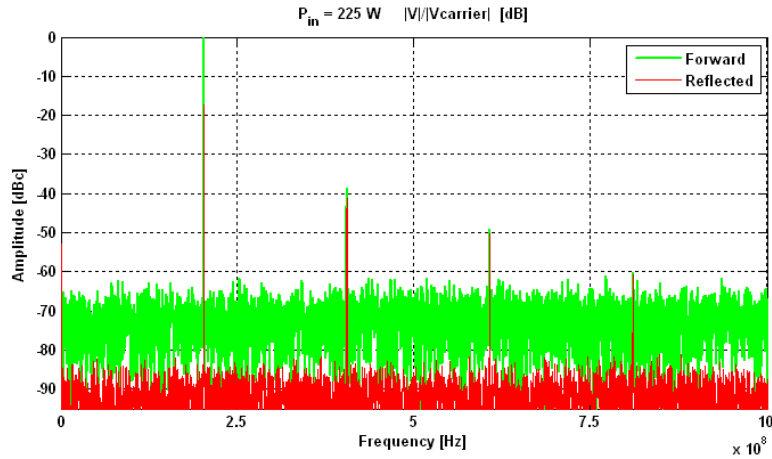


Figure 7.12: Spectrum of incident and reflected voltage for prototype 2.

7.3.3 Conclusion

Although printed inductors offer many advantages – especially in this application – their biggest disadvantage showed clearly: The final component is hard to predict. There are too many factors influencing the inductance value, it is very hard to get it right on the first try. That is the reason, why usually design samples are built and measured, adjusting the geometry in multiple design recursions. This was not possible in the limited available time and neither intended.

Prototype 2 showed well enough – already on its first design recursion – that it is possible to manufacture the rectifier industrially in large numbers with minimum costs, while still obtaining RF/DC efficiencies well over 80%. This further emphasizes the suitability of this rectifier architecture for the energy recovery application.

8. Conclusions and Outlook

8.1 Conclusion

Particle accelerators often dissipate a considerable fraction of their total RF power in dummy loads. The Cern SPS is an example of such a machine and is used in this report, to show different ideas regarding the recovery of that power.

The 200 MHz RF is converted to an intermediate DC voltage, to guarantee the least possible interference with the accelerator system. Solid state diodes were chosen from the range of available technology, to take advantage of recent semiconductor developments, and to avoid the many drawbacks of ultra high vacuum devices.

The RF signal is split to many channels comprising of 1 kW RF power each, as no single diode can withstand the power levels of a SPS termination load. A large array of rectifier modules converts the RF to DC, which will be combined in a single DC channel again. Commercial solar power converters can be used to feed the recovered power back to the utility grid. Each RF/DC module contains four rectifiers – handling 250 W each – and additional components to ensure a graceful degradation of the overall system. If one or more rectifiers fail, the modules will dissipate their power thermally.

Several currently available Schottky diodes were compared in terms of their performance within a rectifier circuit. Maximum power – normalized to the parasitic junction capacitance – was the main criterion for choosing a diode. State of the art devices, built from wide band gap materials, showed the best performance values in that respect. In addition the package and its parasitic inductance play a very important role in RF applications.

Literature on wireless power transmission and high frequency DC/DC converters was evaluated and different rectifier architectures were compared. The most suitable architecture was found to be a series resonant current output rectifier, this configuration exploits the reactive parasitic components of the diode to force a sinusoidal current through the circuit; its operation is closely related to an inverted class E amplifier. The resonant operation naturally restricts the radiation of higher frequency components, which allows the use of a very simple input filter.

Two prototypes of this rectifier were built. Impedance matching was done using series connected transmission lines. The prototypes were evaluated with up to 400 W of pulsed power. RF/DC efficiency, and the complex reflection coefficient were measured for varying input powers. An oscilloscope was used to obtain time domain data of the RF incident and reflected waves. A Matlab script automatically analyses the data, calculates and plots the measurement values of interest. The measurement setup corresponds to an one port, large signal, pulsed, vector network analyzer. In addition it was shown how to derive a nonlinear x-parameter model from the measurements, allowing to simulate the input port behaviour of large rectifier arrays.

The first prototype showed a maximum efficiency of 88.7 % at 284 W. For input powers > 50 W, efficiency is almost constant, regardless of input power level. The prototype proved that high power Schottky diodes can be used at RF frequencies. The overall cost was less than 50 €, making this a very attractive solution for use in large rectifier arrays.

The second prototype was produced with planar inductors, consisting of circuit board traces, improving cooling, further lowering its price, reducing component count as well as making the rectifier circuit more suitable for mass production. The resulting inductances were too big in the first design recursion, nonetheless it showed a maximum efficiency of 88.0 % at 224 W input power.

Both prototypes showed, that the development of a simple, economic, cost efficient module for energy recovery is feasible. With current semiconductors, a realistic power rating of 250 W at RF/DC efficiencies $> 80\%$ can be reached. Moreover, the energy recovery system can be deployed without any impact on accelerator operation.

8.2 Outlook

Simulations indicate that the IDD04SG60C SiC diodes are very promising devices for powerful RF rectifiers. Unfortunately, no samples could be obtained as of yet. Their performance should be evaluated in a RF/DC converter prototype also carrying out continuous wave measurements.

After a complete RF/DC module has been built, the measurement procedure from Chapter 6.2.2 can be used to acquire a nonlinear model of the module. This allows to investigate how an array of RF/DC modules influences the SPS cavities in a simulation environment. Characteristics, such as graceful degradation or harmonic radiation of the entire array can be examined with minimum effort.

The rectifier's power rating could be extended by using SiC or GaN devices with wider band gaps. However these devices operate naturally at higher voltages, resulting in high reactive charging currents into the junction capacitance. Further research can be focused on the investigation of how far **reactive** currents are allowed to exceed the device's forward current ratings, as they do not flow directly through the junction like a DC current.

Semiconductor technology is in continuous development, faster and more powerful devices will be available in the near future. In fact, a prototype of a GaAs Schottky diode was acquired through the "Ferdinand-Braun-Institut für Höchstfrequenztechnik". It was measured, modelled and a rectifier that operates at 2.5 GHz was designed and simulated. It reached 16.4 W with a RF/DC efficiency of 89 %, which is almost three times the power rating of the most powerful rectifier found in literature at that frequency. Those devices are particularly interesting for a rectenna array with very high power density, low weight and size. Revising the idea of wireless power transmission and making it suitable for new applications, like wireless powered sensor networks or home appliances, could be a topic of further research.

Bibliography

- [1] G. Dome. The sps acceleration system travelling wave drift—tube structure for the cern sps. In *Proton Linear Accelerator Conference*. Cern, sep 1976.
- [2] L. Arnaudon, P. Baudrenghien, O. Brunner, and A. Butterworth. Operation experience with the lhc rf system. In *International Particle Accelerator Conference*. Cern, May 2010.
- [3] Eric Montesinos. private communication. Cern, March 2010.
- [4] I. Bustinduy C. Carlile H. Hahn M. Lindroos C. Oyon S. Peggs A. Ponton K. Rathsmann R. Calaga T. Satogata A. Jansson M. Eshraqi, M. Brandin. Conceptual design of the ess linac. International Particle Accelerator Conference, May 2010.
- [5] Fritz Caspers, Michael Betz, Alexej Grudiev, and Hans Sapotta. Feasibility study of a high power RF - rectifier for an energy recovery application. Technical report, CERN, Geneva, May 2010.
- [6] Fritz Caspers. Juas rf course 2010. In *JUAS 2010 proceedings*, Archamps, France, January 2010. Joint Universities Accelerator School, Cern.
- [7] Wikipedia. Solar thermal energy. http://en.wikipedia.org/wiki/Solar_thermal_energy, October 2010.
- [8] Sopogy Hawaii Solar Solutions Center. Power generation application note. <http://www.sopogy.com/solutions/>, October 2010.
- [9] Ian Wilson and Heino Henke. The lep main ring accelerating structure. Geneva, November 1989.
- [10] Vladimir A. Vanke. High power converter of microwaves into dc. Journal of Radioelectronics N9, <http://jre.cplire.ru/iso/sep99/1/text.html>, September 1999.
- [11] William C. Brown. The history of power transmission by radio waves. IEEE Transactions on Microwave Theory and Techniques, VOL. MTT-32, NO. 9, May 1984.
- [12] William C. Brown. The technology and application of free-space power transmission by microwave beam. Proceedings of the IEEE, VOL. 62, NO. 1, January 1974.
- [13] Ronald J. Gutmann. Application of rf circuit design principles to distributed power converters. IEEE transactions on industrial electronics and control instrumentation, vol. ieci-27, no. 3, August 1980.

- [14] F.T. Dickens F.M. Magalhaes W. Strauss W.B. Suiter W.A. Nitz, W.C. Bowman and N.G. Ziesse. A new family of resonant rectifier circuits for high frequency dc-dc converter applications. IEEE, September 1988.
- [15] W.C.Bowman, J.F. Balicki, F.T. Dickens, R.M. Honeycutt, W.A. Nitz, W. Strauss, W.B. Suiter, and N.G. Ziesse. A resonant dc-to-dc converter operating at 22 megahertz. IEEE, September 1988.
- [16] J. M. Borrego R. J. Gutmann D.P. Garrison, A. Lam. Shunt-mounted harmonically-tuned power rectifier for distributed power converters. IEEE, January 1993.
- [17] Riad Samir Wahby. Radio frequency rectifiers for dc-dc power conversion. MIT, May 2004.
- [18] Juan Rivas. Radio frequency dc-dc power conversion. MIT, April 2006.
- [19] Olivia Leitermann Anthony D. Sagneri Yehui Han David J. Perreault Juan M. Rivas, David Jackson. Design considerations for very high frequency dc-dc converters. MIT, June 2006.
- [20] Patrick Marchand, Robert Lopes, Jean Polian, Fernand Ribeiro, and T Ruan. High power (35 kw and 190 kw) 352 mhz solid state amplifiers for synchrotron soleil. page 3 p, 2004.
- [21] W. Schroen, J. Beaudouin, and K. Hubner. Failure mechanisms in high power four-layer diodes. pages 389 –403, sep. 1964. doi: 10.1109/IRPS.1964.362300.
- [22] Jose M. Borrego Ronald J. Gutmann. Power combining in an array of microwave power rectifiers. IEEE Transactions on Microwave Theory and Techniques, VOL. MTT-27, NO. 12, December 1979.
- [23] W. Sinclair H. P. Kindermann, W. Herdrich. The rf power plant of the sps. IEEE Transactions on Nuclear Science, Vol. NS-30, No. 4, August 1983.
- [24] Dariusz Czarkowski Marian K. Kazimierczuk. *Resonant Power Converters*. John Wiley & Sons, Inc., May 1995.
- [25] Inc. Directed Energy. The destructive effects of kelvin leaded packages in high speed, high frequency operation. IXYS, August 1998.
- [26] Dr. Günter Berndes. Is the lowest forward voltage drop of real schottky diodes always the best choice? IXYS Semiconductor GmbH, June 1999.
- [27] Marian K. Kazimierczuk. Analysis of class e zero-voltage-switching rectifier. IEEE transactions on circuits and systems, VOL. 37, NO. 6, June 1990.
- [28] Marian K. Kazimierczuk. Class e resonant rectifier with a series capacitor. IEEE transactions on circuits and systems-I fundamental theory and applications, VOL. 41, December 1994.
- [29] John S. Shafran David J. Perreault Juan M. Rivas, Riad S. Wahby. New architectures for radio-frequency dc–dc power conversion. IEEE transactions on power electronics, VOL. 21, NO. 2, March 2006.

- [30] Keiichi Sakuma and Hirotaka Koizumi. Influence of junction capacitance of switching devices on class e rectifier. Tokyo University of Science, sep 2009.
- [31] Prof. Dr.-Ing. Hans A. Sapotta. Skript zur Vorlesung Hochfrequenztechnik. Hochschule Karlsruhe, June 2006.
- [32] Hans-Joachim Würfl. private communication. Berlin Microwave Technologies AG, March 2006.
- [33] Robert F. Pierret. Semiconductor device fundamentals. Addison Wesley, April 1996.
- [34] Ian Getreu. Modeling the bipolar transistor. Elsevier Science Ltd, August 1978.
- [35] Holger Kapels Michael Krach Ilia Zverev Christian Miesner, Roland Rupp. thinq!TM silicon carbide schottky diodes: An smps circuit designer's dream comes true! <http://www.infineon.com>, February 1999.
- [36] Kent Walters. Rectifier reverse switching performance. Microsemi application note, August 1998.
- [37] L. M. Tolbert B. Ozpineci. Comparison of wide-bandgap semiconductors for power electronics applications. Oak Ridge National Laboratory, <http://www.osti.gov/bridge/>, December 2003.
- [38] Jon Schleisner. Design guidelines for schottky rectifiers. Vishay General Semiconductor, August 2008.
- [39] Helmut Föll. Semiconductors i. http://www.tf.uni-kiel.de/matwis/amat/semi_en/, August 2008.
- [40] B. Janke C. Meliani W. Heinrich J. Würfl P. Kurpas, A. Wentzel. Monolithically integrated gainp/gaas high-voltage hbts and fast power schottky diodes for switch-mode amplifiers, May 2009.
- [41] Julius O. Smith III. *Introduction to Digital Filters: with Audio Applications*. W3K Publishing, October 2007. http://www.dsprelated.com/dspbooks/filters/Zero_Phase_Filters_Even_Impulse.html.
- [42] Michael Hiebel. *Fundamentals of Vector Network Analysis*. Rohde & Schwarz, Muenchen, Germany, 2008.
- [43] David E. Root Jan Verspecht. Polyharmonic distortion modelling. IEEE microwave magazine, June 2006.
- [44] Sophocles J. Orfanidis. *Electromagnetic Waves and Antennas*. Rutgers University, www.ece.rutgers.edu/~orfanidi/ewa, 2008.
- [45] Stephen P. Boyd Sunderarajan S. Mohan, Maria del Mar Hershenson and Thomas H. Lee. Simple accurate expressions for planar spiral inductances. IEEE journal of solid-state circuits, VOL. 34, NO. 10, October 1999.

Contents of the CD-Rom

Path	Description
./Eagle_PCB/	Printed circuit board layouts of prototype 2. The files are readable by the program “Eagle”.
./Latex/	Latex source code and all figures for this thesis and the IPAC publication.
./Literature/	Collection of all the literature, referenced to in this work.
./Matlab_Scripts/	Matlab source code for deriving efficiency and complex reflection coefficient from a set of .isf files (Tektronix data file format). Those scripts realize the measurement methods described in Chapter 6.2.
./Measurements/	Results from the measurements of the two prototypes as image files. The oscilloscope data files can be found in <i>raw.zip</i> .
./PowerPoint/	Various power point slides, presented at Cern section meetings throughout the duration of this work.
./Rectifier_Comparison/	Excel table with the results of the comparison of different rectifier architectures.
./Rectifier_Comparison/PSpice	The PSpice projects for the simulations, the rectifier comparison was based on.