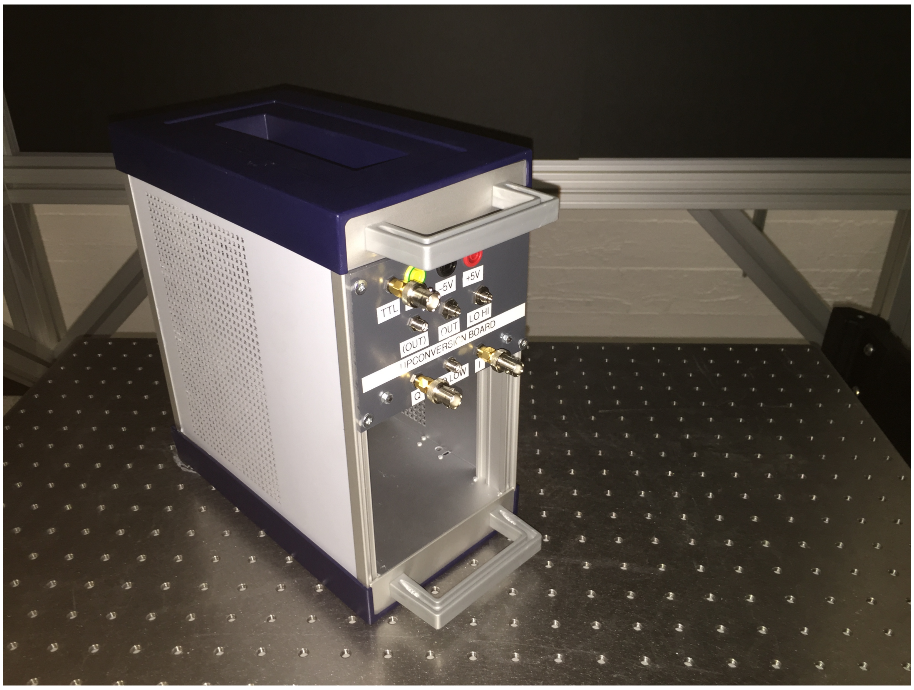
ETH Zurich

Semester thesis

Up conversion board for amplitude and phase controlled microwave pulses on the 3-26.5 GHz scale



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Abstract

The control of phase and amplitude of pulses in the K band is not common practice, as the decreasing wavelength asks for higher demands of the microwave components. This work documents the implementation of an Up conversion of controlled pulses from 500 MHz to the 3-26.5 GHz range, using an electrical circuit with standard microwave components. Typical measurements show, that a very clean spectrum with 60 dBc and output power of -24 dBm (tested up to -11 dBm) for arbitrary pulse shapes is feasible.

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1 Introduction

1.1 Rydberg atoms and the need for high microwave frequencies

Any atom with one electron being excited to a high principal quantum number 'n' is said to be a 'Rydberg atom'. Such atoms have exaggerated properties regarding the interaction with electromagnetic fields, which will be briefly covered in this section. Their field of study becomes increasingly important with the advances in cavity quantum electrodynamics [1]. Because of the high radial quantum number 'n', the excited state of the atom can be approximated by the model of an electron rotating around a core. The electron being far away from the center, it is object to only the net charge (electrons + protons) of the core. Therefore any such atom in a Rydberg state can be treated as a special version of a hydrogen atom. Solving the hydrogen atom means solving a two-body problem. Within this model it feels naturally to conclude from the large radial separation of electron to core, that Rydberg atoms can have large dipole moments $d \propto \langle r \rangle \propto n^2$. Since the coupling of an atom to a field mode is given by $g = \frac{d \cdot E}{\hbar}$, larger dipole d means larger coupling.
The atomic system can be well described by three quantum numbers (n, l, m) associated with the outer electron. When restricting ourselves to two distinct states on the m-manifold (e.g. 37p and 37s with $m = 0$), one can refer to the atom as a 'qubit' with the two states labeled $|0\rangle$ and $|1\rangle$. Such quantum systems are well understood. In a rigorous treatment of the interaction between atom and electromagnetic field it can be derived how the system can be manipulated by applying various electromagnetic pulses with certain amplitude and frequencies. For this thesis helium Rydberg atoms in states (37,s) as $|0\rangle$ and (37,p) as $|1\rangle$ are used. The energy difference between these two states corresponds to the energy quantum with frequency of approximately 25 GHz. Generation of electromagnetic pulses in the frequency range of 20-30 GHz is therefore needed to obtain full control of the quantum system.

1.2 Rydberg experiment at ETH Zurich

In the following the setup in which context this thesis was done will be explained shortly. Pulsed supersonic helium clouds are coming out of the valve with an average speed of 1.7 km/s at a rate of 25 Hz. They are excited to the metastable $(1s^1)(2s^1)\,{}^1S_0$ state by an electric discharge. A skimmer then acts as a transverse velocity filter to only keep the coldest atoms in the cloud, as the warm part diverges from the beam axis. Two razorblades allow to adjust the shape of the atom beam. As soon as the atoms reach zone 1, they will be transferred to the ns state, usually (37,s), by a frequency doubled YAG laser at a wavelength of $\lambda = 313\,\mathrm{nm}$ ($R = 25\,Hz, \Delta t = 10\,\mathrm{ns}$) and a subsequent microwave pulse ($\Delta t = 160\,\mathrm{s}$). Zone 2 is the interaction zone, where microwave pulses can be applied either by using a horn antenna or by using a transmission line mounted on the sample holder. These transfer part of the population from the (37,s) state to the (37,p) state. As the atoms enter zone 3, the (37,p) state will have decayed radiatively, but not the (37,s) state. This is due to the very short lifetime of the (37,p) state, compared to the traveling time between zones 2 and 3. By applying an ionization voltage of up to 128 V for the s-state between electrodes 4 and 5, the Rydberg atoms are ionized which can be considered as a quantum mechanical measurement process. Any atom in the (37,s) state will yield an electron, which is then accelerated by the applied voltage towards an micro channel plate (MCP) detector. The resulting image of measured electrons yields an ensemble averaged measurement of the Rydberg atom in the $|0\rangle, |1\rangle$ basis, as the lack of an electron can be understood as the (37,p) state. For a more detailed description see [2].

The following sections focus on how to generate amplitude and phase controlled microwave pulses, which can then be used to manipulate the system in zone 2.

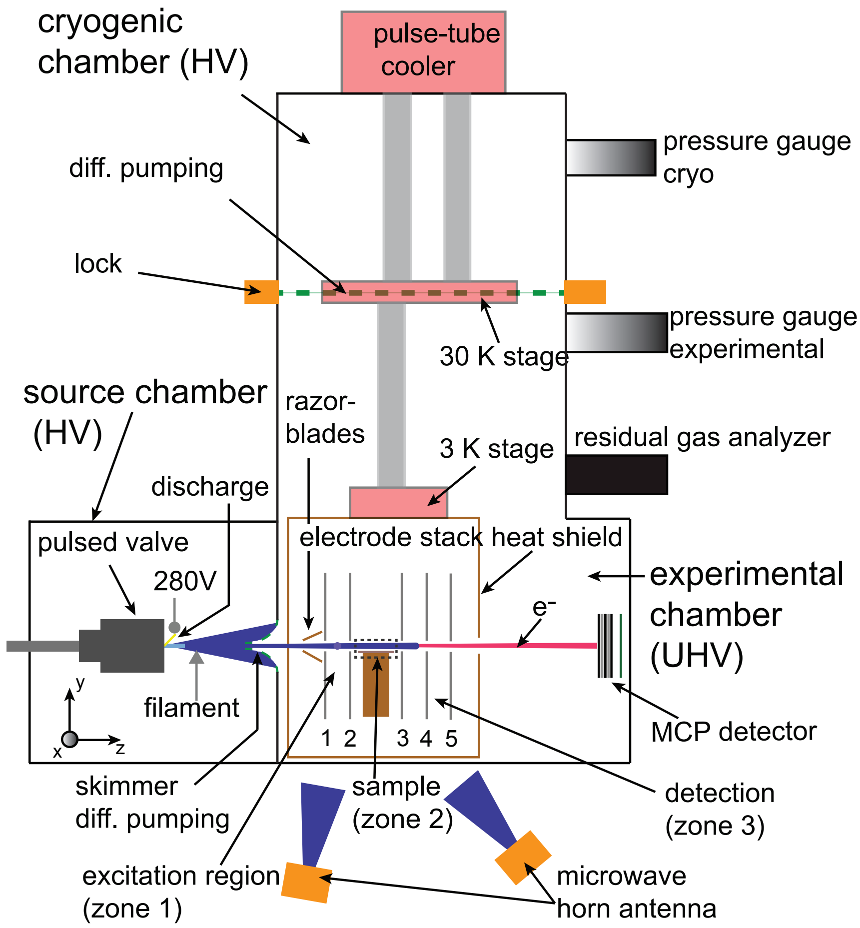

Figure 1: Schematic overview over the Rydberg experiment setup at the ETH Rydberg lab.

2 Pulse generation and up conversion

This section describes the theory for a method to generate phase and amplitude controlled waveforms at frequencies in the 3-26.5 GHz range using an 'Arbitrary Waveform Generator' and an electronic circuit, called 'up conversion board'.

2.1 Arbitrary Waveform Generator (AWG)

The AWG is a commercial device, which is able to realize wave patterns in the 0-500 MHz range. The version used in the Rydberg lab is a Textronix AWG5014C. Arbitrary pulse patterns can be loaded onto the device. Four BNC output ports exist with adjustabe amplitude and phase delay, as well as the possibility to add a DC offset bias via an extra input at the back. Another four ports for DC output voltages exist.
There already exist mathematica files, which produce pattern files that can easily be uploaded to the generator[1].

2.2 Theory of mixing

The main task is to convert the 500 MHz waveform to a higher frequency. A standard approach is a method called 'mixing' and can be compared to the subject of nonlinear optics in a Kerr-active medium, where the nonlinear terms of the Kerr nonlinearity allow for an interaction between two frequency components. Needed is an element, which takes two inputs and has their multiplication as an output. Mixing in electrical engineering can be achieved using a 'diode'. A sinusoidal tone at frequency f_1, usually called 'Local Oscillator' (LO), enters the diode. The transmittance of this signal through the diode depends on the voltage applied to the diode at a port called 'IF', for 'Intermediate Frequency'. When the 'IF' signal oscillates with some frequency f_2 (most commonly f_2 is a lower frequency compared to f_1), the 'LO' signal will thus be modulated by this frequency. Ideally, the output signal contains only the so called 'second order', namely the simple multiplication of the two frequencies. As shown in the next chapter, the output of such an ideal mixing process should contains the superposition of two waves with frequencies $|f_1 - f_2|$ and $|f_1 + f_2|$. These two frequencies are called 'left' and 'right' sideband. As a real device can hardly be controlled to only behave in this desired way, the output will always also contain other frequencies called 'higher harmonics', as well as a significant part of the original f_1 signal leaking through.
As one normally is only interested in a single output frequency, ways need to be found to avoid or filter spurious sideabands away. This will be subject of the next section. More details on 'mixing' can be found in [3].

[1] A library for arbitrary gauss or square pulses can be found here: R/USERS/Joannis/PulseGen

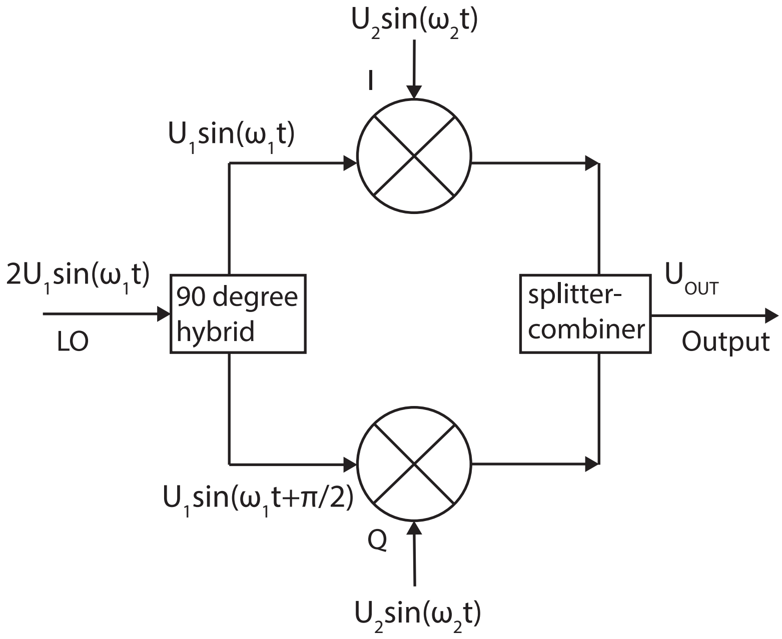

Figure 2: Circuit diagram for an In-Phase/Quadrature mixer.

2.3 IQ-mixer

One elegant way to suppress the the output of one of the two sidebands is using the electronic circuit shown in Figure 2, also called In-phase/quadrature mixer, or abbreviated 'IQ-mixer'. The circuit can be understood in the following sample calculation. Consider the sinusoidal input 'LO' given by $U(t) = 2\,U_1 sin(\omega_1 t)$. This signal is equally split into two channels, one of them introducing a $\pi/2$ phase shift. The signals in the respective channels are thus $U_1 sin(\omega_1 t)$ for I and $U_1 sin(\omega_1 t + \pi/2)$ for Q. Now let 'I' be subject to a mixing process (signal multiplication) with a signal $U_2 sin(\omega_2 t)$ and Q with a signal $U_2 sin(\omega_2 t + \varphi)$, simply shifted by a phase φ. Eventually, both waves are recombined (superposed) to yield an output signal:

$$\begin{aligned} U_{out} &= U_1 U_2 \big[sin(\omega_1 t)\, sin(\omega_2 t + \varphi) + cos(\omega_1 t)\, sin(\omega_2 t)\big] \\ &= \frac{U_1 U_2}{2} \big[cos\big((\omega_1 - \omega_2)t - \varphi\big) + cos\big((\omega_1 + \omega_2)t + \varphi\big) - sin\big((\omega_1 - \omega_2)t\big) + sin\big((\omega_1 + \omega_2)t\big)\big]. \end{aligned}$$

There are two particularly neat cases. For $\varphi = \pi/2$ the right sideband and for $\varphi = 3\pi/2$ the left sideband vanishes due to interference. Therefore an IQ-mixer gives the possiblity to perform a mixing process with a sideband removed.

The problem of the initial frequency ω_1 still appearing in the output comes from the fact, that DC voltage offsets exist within the mixer and bias the built in diode. Application of very small DC voltage offsets to the I and Q signal, this effect can be compensated. Further information on IQ-mixers in general are described in [4].

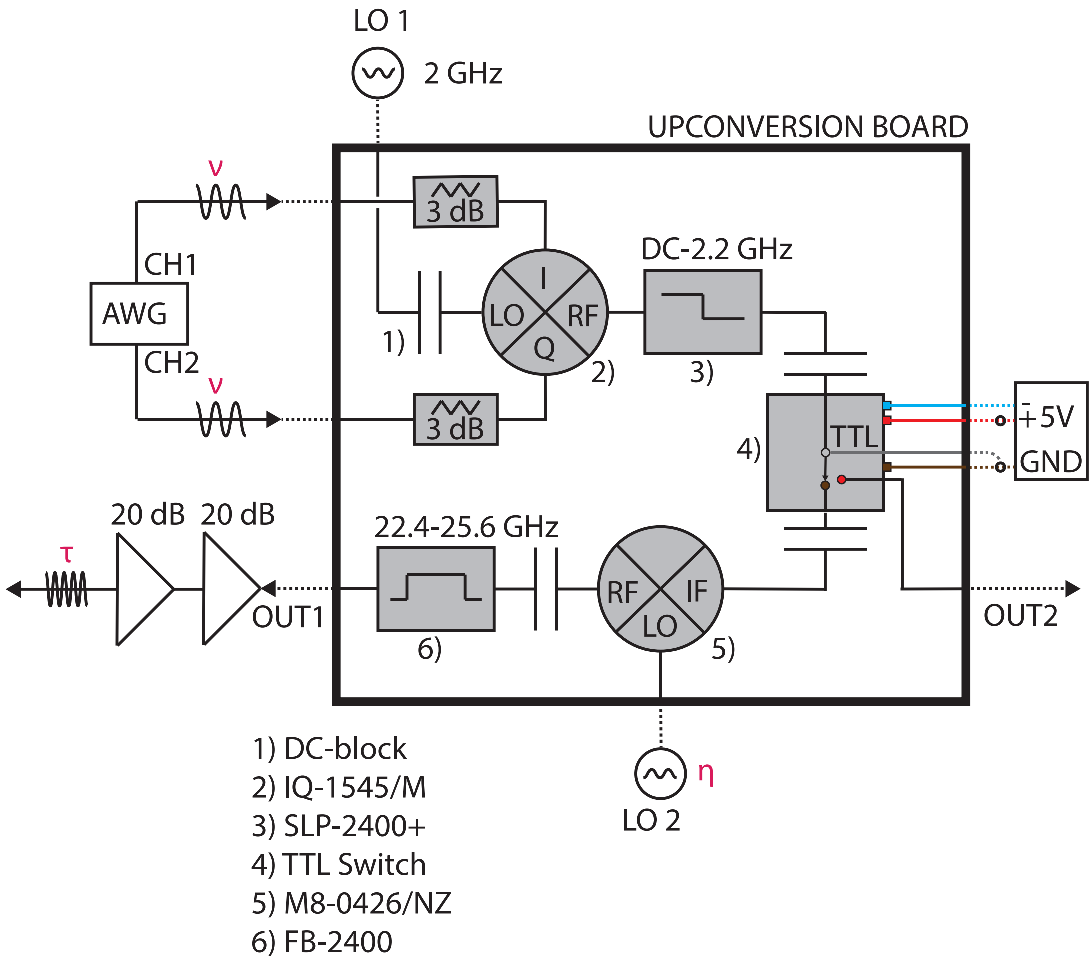

Figure 3: Circuit diagram of a setup realizing programmed pulses at a 3 to 26.5 GHz scale using an up conversion board. All dotted lines represent external (not onboard) connecting cables. The output port can be switched from 'Out1' to 'Out2' by changing the connection of the TTL signal from 'GND' to '+5V'.

2.4 Up conversion board

As was shown in the previous section, the basic principles of mixing already allow one to think of a straightforward implementation to take an arbitrary pulse shape at the MHz range and use a mixer and another oscillating signal at the GHz range to get an output containing the pulse shape at the desired frequency. The main motivation to use a more complex setup to do this job is on the one hand the lack of suitable IQ-mixers in the k-band (18-26.5 GHz), on the other hand to increase the carrier to sideband distance, such that spurious sidebands are far detuned from the atomic resonance. At the same time one has to try that the final output power of the signal does not get too small. A similar method to the one described in the proceeding chapter was realized in [5]

The easiest way to remove higher harmonics or sidebands is to use a bandpass-filter with only frequencies passing in the region of the target frequency τ. Since filters with very narrow bands are expensive, this frequency noise need to be separated by several GHz from τ to allow good filtering. This will be the case, if the modulating signal (the one with the lower frequency), usually called 'IF', is already in the GHz range for the mixing process. Therefore, a preceding intermediate up conversion is needed. For the low GHz regime it is possible to use an IQ-mixer for this job. This also gives the possibility to

intermediately create a clean up converted signal at a few GHz.
The proposed arrangement of components shown in Figure 3 aims to mix an arbitrary waveform coming from the AWG at frequency ν (DC-500 MHz) to a frequency τ (3-26.5 GHz) and at the same time to create a clean output spectrum. Two identically shaped pulses are generated at a frequency ν via the AWG and used as the signals 'I' and 'Q' for the IQ-mixer. Before entering the mixer, both signals go through a 3 dB attenuator to avoid accidentally driving with too much power, as well as to improve impedance matching. A microwave tone at 2 GHz comes from a local oscillator ('LO 1') and is modulated by the 'I' and 'Q' via the mixer. The phase and amplitude of the I- and Q-pulse are typically adjusted to cancel the left sideband. Higher harmonics occurring at multiples of 2 GHz are filtered out by a lowpass filter with a passband from DC-2.2 GHz. The dominating frequency of the output thus is the right sideband at $(2+\nu)$ GHz. An installed switch gives the possibility to monitor the output of the mixer at 'Out 2'. Typically, the signal is used as the new 'IF' input for a double balanced mixer. A second up conversion takes place, using the sinusoidal signal of a second oscillator ('LO 2') at a particular frequency η in the GHz range (20-26.5). The desired frequency τ can be interpreted to be the left $(-)$ or right $(+)$ sideband of this mixing process with frequency

$$\tau = \eta \pm (2 + \nu).$$

Choosing the left sideband $(-)$ to yield the desired τ is motivated by the bandpass filter with a bandpass from approx. 22 to 26 GHz installed after the second mixer. For the Rydberg experiment, τ will be very close to 26 GHz. If this corresponds to the left sideband, the right sideband will be located at $26 + 2(2 + \nu)$GHz. This is deep in the highly damped region of the filter. Hence, the right sideband will be absorbed and we are left with only one sideband. The final output can optionally be magnified by two external 20dB amplifiers.
A subtle point to mention is the installment of DC-blocks. As different devices with non common grounds are connected to the board, the danger of unwanted currents flowing along elements needs to be avoided. DC-blocks being placed at the points shown in Figure 3 ensure that each element is connected to only one single ground.

2.5 Connecting components - Impedance matching

All devices within a microwave circuit are typically manufactured with standardized input and output ports. These can be connected using coaxial cables with one of the two common connector-types SMA or SK. Each point in the circuit diagram, where such a port connection has to be established, might add a disturbance in the propagation of the microwave. Whereas the propagation along the coaxial cable is just the usual propagation in a linear medium, a port-connection corresponds to the well-known situation of an electromagnetic field traveling across a surface between two media. This problem is displayed in Figure 4 in the most general case. For a port-connection this simplifies to the case with $\theta_1 = \theta_2 = \theta_3 = 0°$ and no difference between orthogonal or transverse polarization. The strength of transmitted and reflected field, relative to the incoming electric field is given by

$$\frac{E_3}{E_1} = r = \frac{Z_2 - Z_1}{Z_2 + Z_1} \tag{1}$$

$$\frac{E_2}{E_1} = t = 1 + \frac{Z_2 - Z_1}{Z_2 + Z_1}, \tag{2}$$

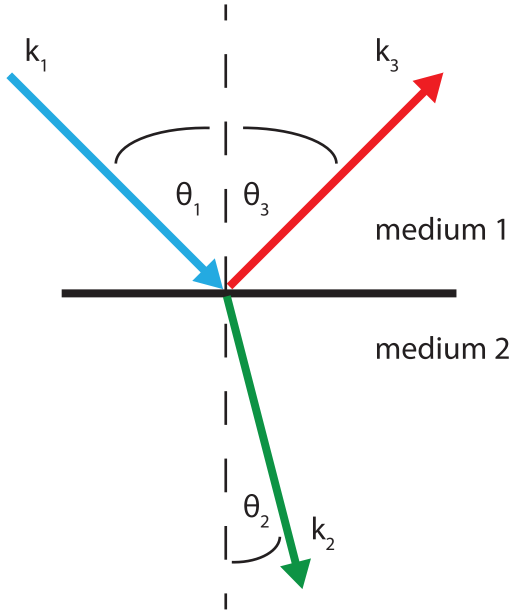

Figure 4: Incoming electric field hitting a surface between two media. k_i denotes the corresponding wavevectors of incident, transmitted and reflected wave.

where Z_i is the so called impedance of medium i and given by

$$Z_i = \sqrt{\frac{\mu_0 \mu}{\epsilon_0 \epsilon}}.$$

If the impedance across the connector varies at some point, a certain amount of the electric field will be reflected. A canonical example would e.g. be air possibly enclosed between the two surfaces of each port. Z_0 of air usually differs from the impedance of the isolation in the coaxial cable. This is the major problem of these connectors, as standing waves inside elements can arise. Therefore, when working with different components it is very important to have all impedances matched to each other. The standard impedance used in microwave electronics is $Z = 50\,\Omega$. The reflection and transmission characteristics of a component can be measured by a 'Vector Network Analyzer' (VNA). This device is connected to two ports of a component. It measures the reflectivity and transmission coefficients 'r' and 't' for a certain frequency range from each port to the other. The results are denoted 'S11', 'S12', etc. and displayed in a matrix like arrangement. This is also referred to as the 'scattering matrix' [6]. How to manufacture SK-connectors and how to determine their quality is subject of the next chapter.

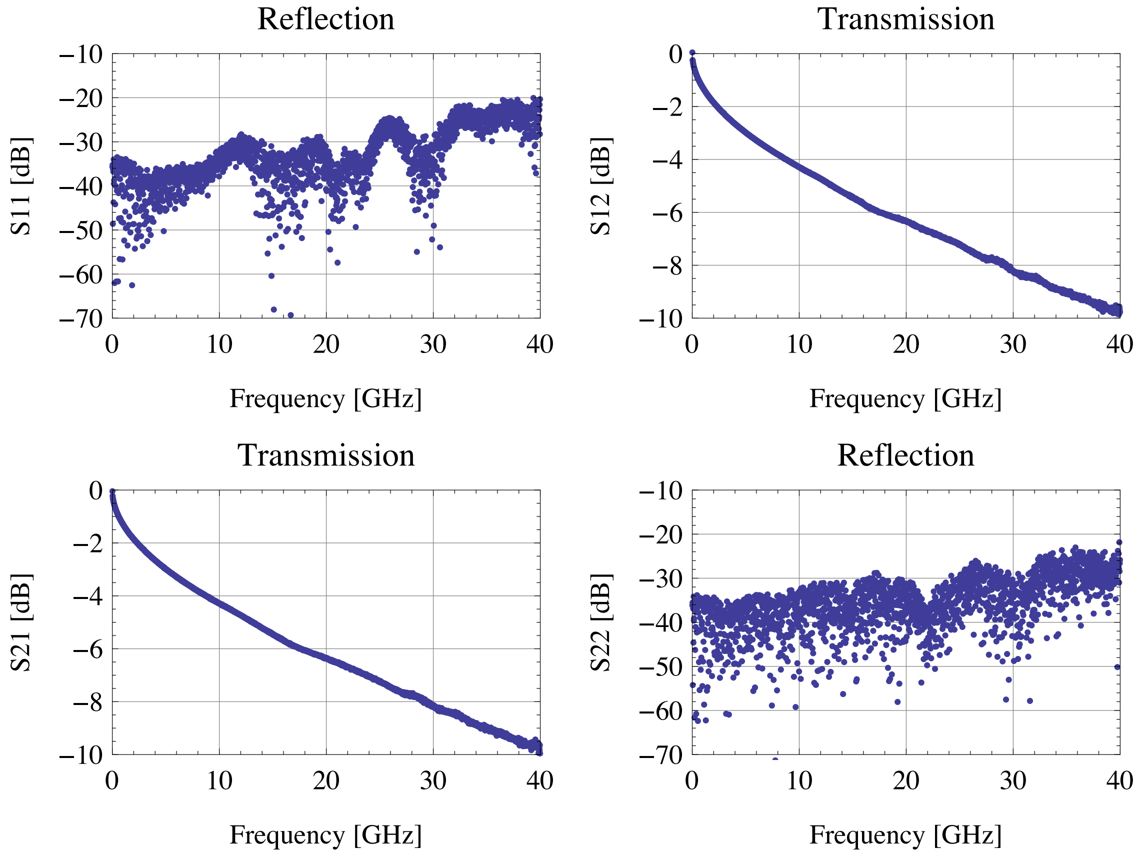

Figure 5: Scattering matrix of an approximately 1.7 m coaxial cable with well manifactured SK-connectors.

3 Assembling the board

3.1 Manifacturing SK-connectors

An important point during the assembly is the guidance of microwaves from one microwave component to the other. The general way to do this is to use coaxial cables. The cable itself can be purchased[2], but the connectors at each end have to be added manually. There are two common types of connectors. SMA being the cheaper one used for frequencies between DC to 20 GHz and SK used for DC to 40 GHz. The theoretical difference is the more precise impedance matching for the latter, allowing higher frequencies to propagate without reflection. For this board only the SK version was used. The quality of a cable can be determined by investigating its scattering matrix using a VNA. A good example is shown in Figure 5. The reflection never exceeds -20 dB, which is a standard requirement, and wiggles in the transmission only show up at very high frequencies with an amplitude of less than 0.2 dB. A list with remarks on the most important manufacturing steps can be found below.

[2]you can find them in HPF D13

1. Cut the outer layer of copper using the little cutting block. It's cutting depth should be adjusted to just not cut the dielectric. To finally remove outer layer, use tweezer to rotate layer around cable axis. Then pull without rotating.

2. Use microscope, box cutter and measuring stick to cut the length to 2.4 mm. If isolation is contaminated, clean with isopropanol and wait for it to dry.

3. When soldering the first element, use a toothpick to distribute flux around the region of interest. It is very important to minimize heat exposure. The isolation will expand and complicate the achievement of good impedance matching. Immediately cool after soldering by applying isopropanol to all heated regions. Do not release the setup. Wait for it to cool down and dry.

4. Use microscope and box cutter to cut and remove the isolation. This is the most crucial part of the process. Do not hesitate to clean box cutter with isopropanol and to use a (new) sharp tip. Strongly press the blade against the connector side, aim such that you will just not hit the center conductor and cut straight down. Then turn the cable and repeat three to four times until isolation can be removed with tweezer from center conductor. The quality of the cut very strongly influences the impedance matching of the connector. Try to achieve a flushed cutt-off with no fluff of isolation left over.

5. When soldering the pin onto the center conductor, cut a very small piece of soldering-tin (≈ 0.5 mm) and use tweezer to place it inside the pin. Position distance plate W262 on the center conductor. Use electrical tweezer to solder pin. If succeeded, W262 should be trapped just hard enough such that one can only remove it with some effort. If the little gap between pin and isolation is contaminated, use compressed nitrogen to clean.

6. Use two wrenches (5,5 and 6) and medium amount of force to fasten outer connector part. If this is done too weakly, one can observe resonances at the VNA. Avoid applying to much force either.

Figure 6: Front panel of the upconvesion board. 'LO LOW' and 'LO HI' denote 'LO 1' and 'LO 2' respectively. 'Out' and '(Out)' correspond to 'Out 1' and 'Out 2'. All other labels are identical to Figure 3

3.2 Input/output specifiactions

Given the specifications of every component in the circuit, a similar statement can be made for the board as a single component. All data are chosen with respect to the 1 dB compression point.

Input Device	Frequency	effective Power
LO 1	$\zeta \equiv$ 1.5-2 GHz	13-16 dBm
LO 2	$\eta \equiv$ 3-26.5 GHz	13-16 dBm
AWG	$\nu \equiv$ DC-500 MHz	up to 9 dBm/ 1.2 Vpp

The effective power of each device denotes the amount that enters the board. When dealing with long connecting cables, attenuation needs to be compensated. The sum $\zeta + \nu$ has to be roughly $\leq 2\,\text{GHz}$ when choosing the right IQ-sideband.

3.3 Integration into the Rydberg setup

To avoid any electrical disturbances on the up conversion process, the whole circuit is fixed on a PVC plate and put into a plastic box. An input/output panel, shown in Figure 6, allows for an easy way to interact with the whole board as a single component. The box is placed close to the experiment. Both LO's are connected with approximately 2 m coaxial cables. Power sources for switch and amplifiers are placed behind the board. The AWG is located next to the LO's and connected by a concatenation of BNC cables, see Figure 7.

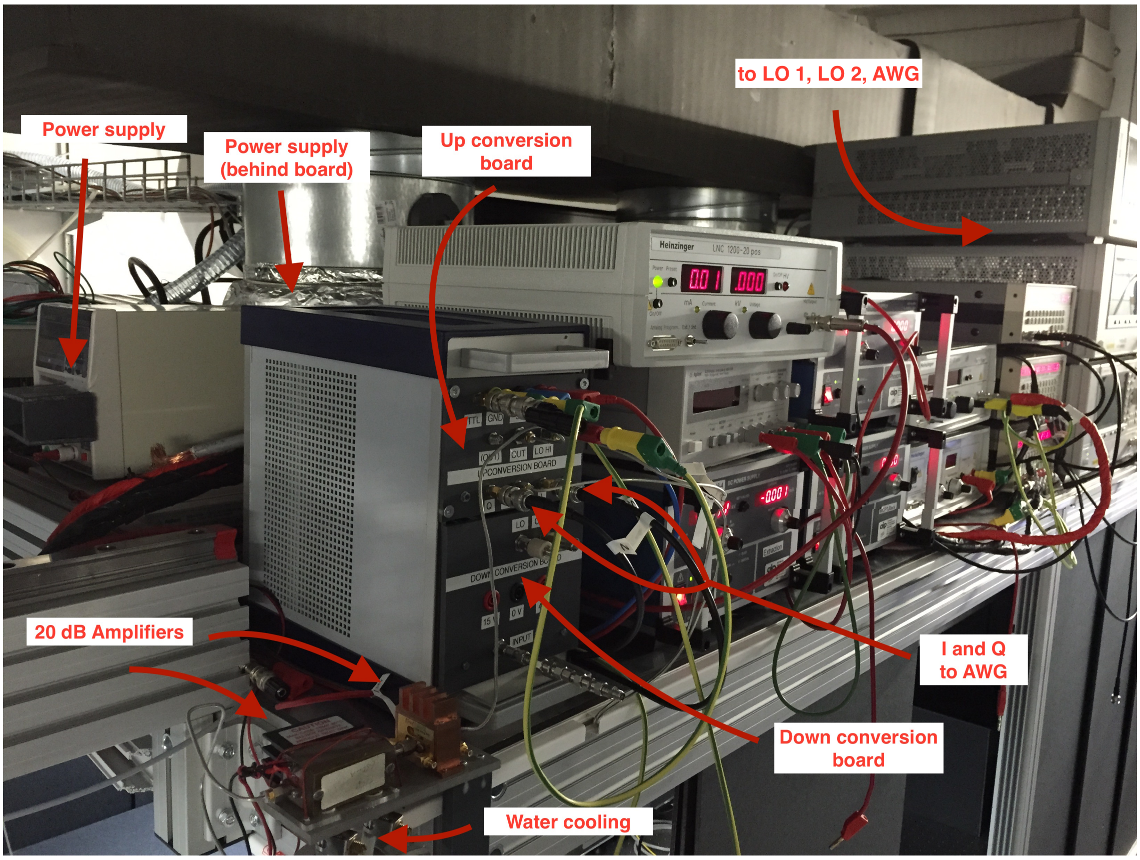

Figure 7: Up conversion board integrated into the experiment.

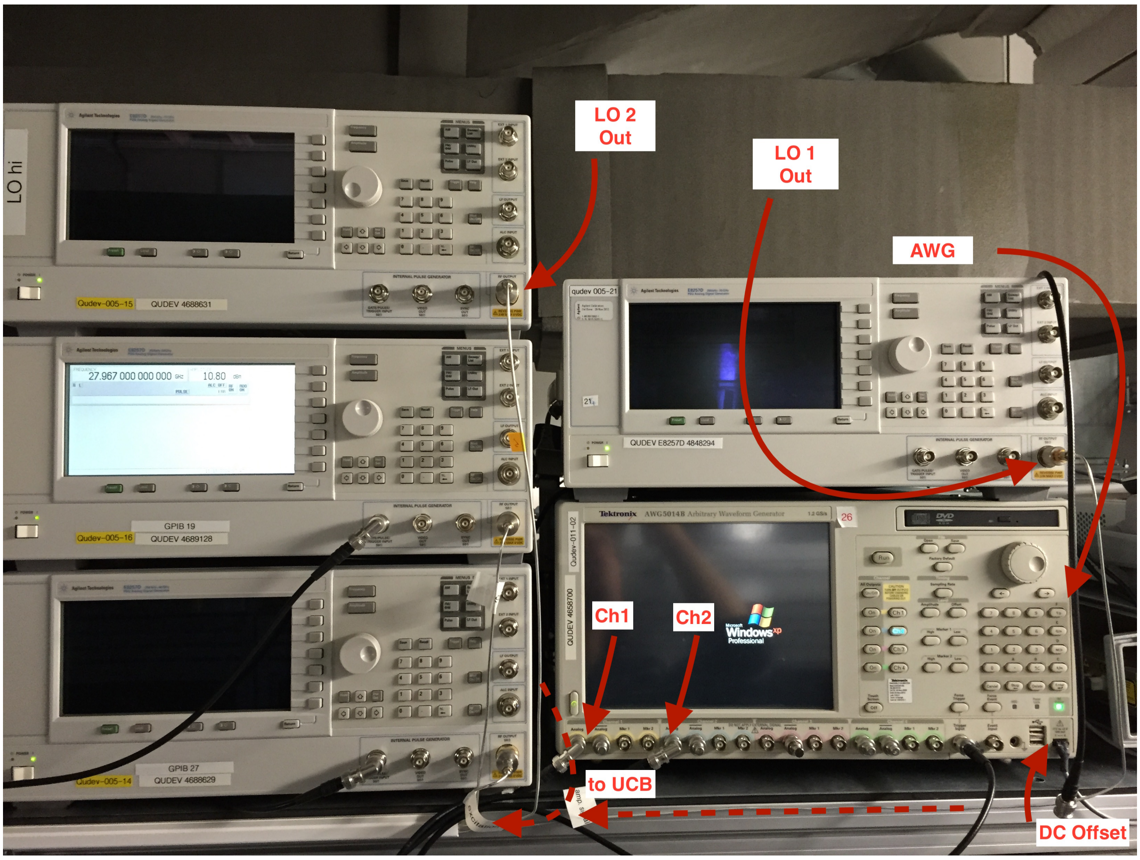

Figure 8: Local oscillators and AWG connected to the Up conversion board.

4 Characterization of the downconversion board

The idea of this section is to measure, if the final up conversion circuit works as it should be. Powerlosses are tracked and understood, IQ-calibration settings are found, the overall spectrum of the process is investigated and finally several pulse shapes are tested and measured using an FPGA (vertex 4) as well as the 'cleansweep' measuring routine of the 'Quantum Device Lab'.

4.1 Calibration IQ-mixer

As shown before, the input of the IQ-mixer can be adjusted in a way, that one sideband vanishes at the output. The according settings for both amplitude and phase have to be determined for each of the channels I and Q. These can be found by a straightforward procedure. LO 1 is set to 2 GHz at a power of 16 dBm. Channels 1 and 2 of the AWG are set to emit a sine wave in continuous mode with frequency and working power close to the desired pulse shape. The TTL switch is set to transmit the output to 'OUT 2' ('GND'), which is connected to a spectrum analyzer. Starting from the theoretical value for the phase difference of e.g. $\varphi = \pi/2$, the phase and amplitudes of both pulses are varied until the power of the undesired sideband is minimized. Typical values for amplitude, phase shift and DC offset are (0.2 Vpp, 0 deg, -1.07 V) and (0.198 Vpp, 30.7 deg, -0.44 V) for Channels 1 and 2 respectively. The DC offsets are reduced by the voltage divider by a factor of 100 before added to the signal.
Another important point is to minimize the leakage from the LO input. Usually, when driving LO 1 with 2 GHz, one would observe a significant amount of a 2 GHz signal at the output. The origin for this are small offsets in the diodes used in the mixing process. The LO input goes through the diode, which has a transmittance depending on the applied voltage coming from the IF input. Only when zero voltage is applied, no signal should come out. A small voltage disturbance thus leads to a leakage of the LO frequency. In order to solve this problem, a small offset DC voltage (mV) can be applied to each of the channels, canceling out the disturbance. This can be done using the DC output channel of the AWG and manipulating it by a voltage divider. The voltage then reenters the AWG via the 'add input' port at the back. This basically follows the standard procedure for applying offset voltage described in [7]. The minimization of the LO 1 leakage can then be done by tuning the DC output at the AWG. An example for a good calibration is displayed in Figure 9. From 9a), it can be seen that a contrast of at least 60 dB from signal to LO and sideband noise is feasible. Part b) of the measurement shows this contrast maintains after the whole up conversion process. Therefore the calibration settings can in general be found, by tuning the parameters and looking at the signal using port 'Out 2', and the experimental setup does not need to be deconnected for finding the parametes.

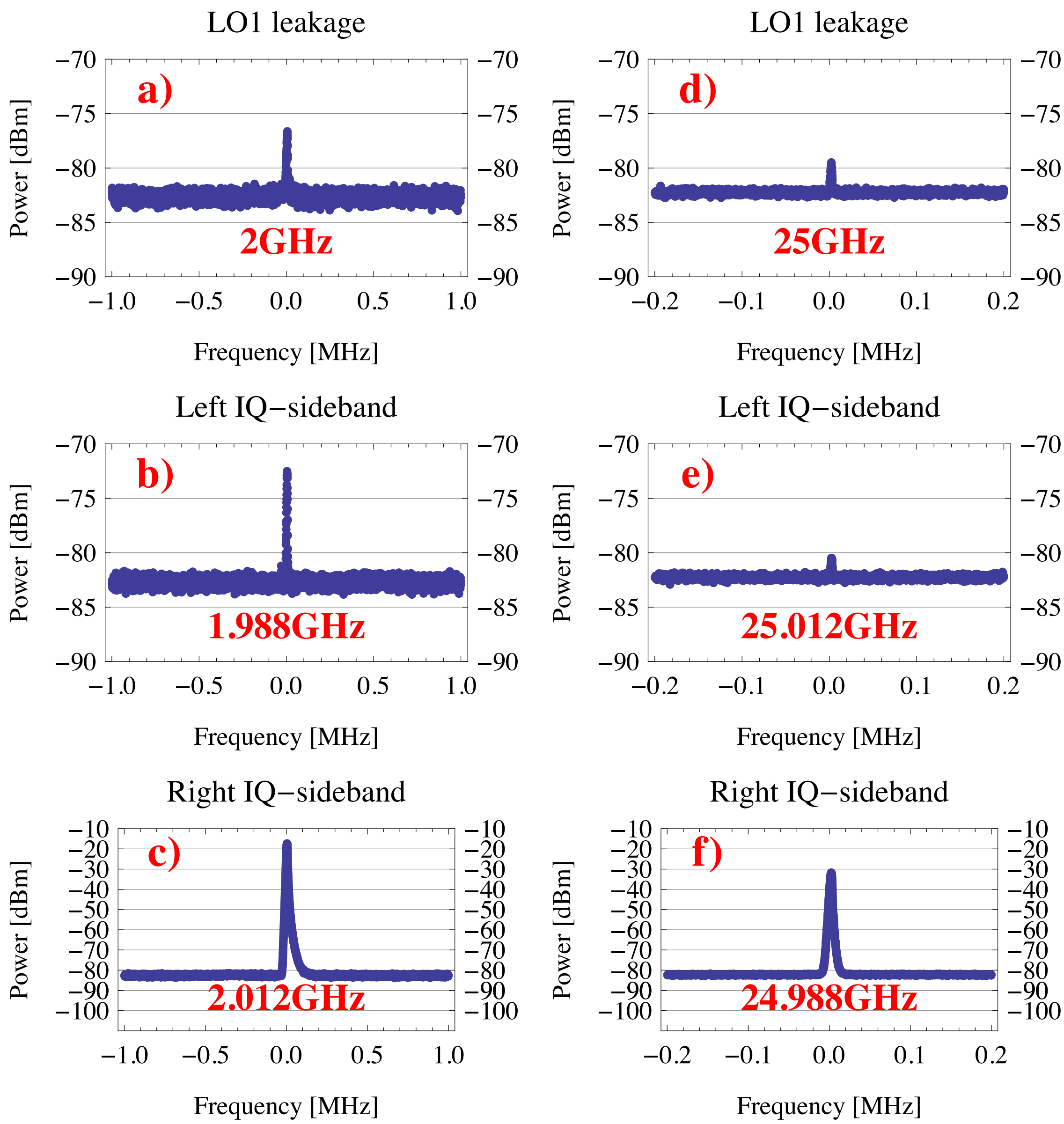

Figure 9: Calibration for a 12 MHz upconverted signal, minimizing left IQ-sideband and LO1 leakage. AWG settings (Amplitude, phase shift, DC offset) for both channels found to be: Ch1(0.2 Vpp, 0, -1.07 V), Ch2(0.198 Vpp, 30.7, -0.44 V). LO1 and LO2 were set to 2 GHz at 19 dBm and 27 GHz at 25 dBm respectively. Signals measured at Out1 and Out2: a-c) Power measured at port 'Out 2' after first up conversion for LO leak, left and right sideband. b) Fraction mapped to the left M8-sideband after whole process, coming out of port 'Out 1' for LO leak, left and right IQ-sideband.

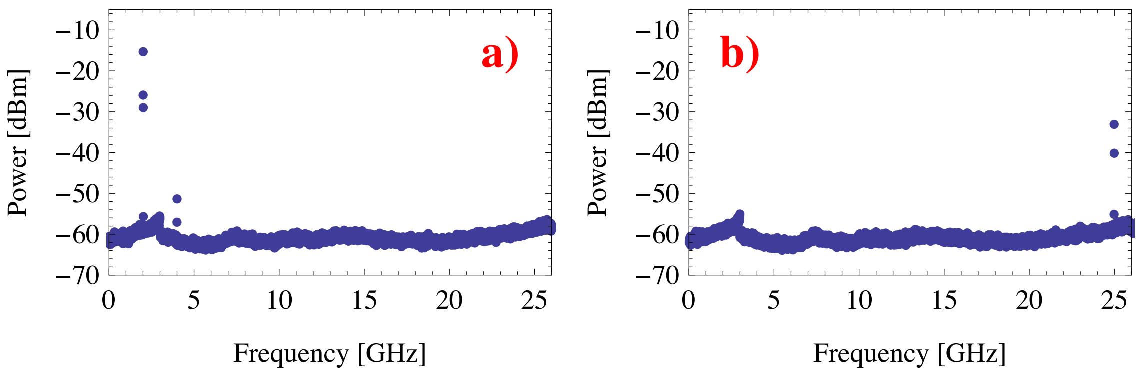

Figure 10: Spectrum of a 12 MHz signal upconverted by 2 GHz and subsequently 27 GHz. a) After first up conversion, measured at port 'Out 2'. b) After whole process, measured at port 'Out 1'.

4.2 Spectrum of up conversion

The resulting output spectrum for a standard sinusoidal input of a particular frequency can be similarly investigated on the two stages 'Out 1' and 'Out 2'. Figure 10 shows a broad overview of all relevant contained frequencies in the signal. No other frequency contribution above the measured noise level of -60 dBm in the final output is an important attribute of the up conversion board. Initially one would expect at least some higher harmonics lying above the noise level. Both filters used in the circuit significantly contribute to avoid this effect. The almost similarly clean spectrum after the first up conversion, see Figure 10a), is an important prerequisite for a clean second mixing process. Therefore it can be stated, that for an up conversion with settings as in Figure 10 there are no other frequency contributions in the intervall [0, 26.5 GHz] up to a range of at least 30 dB from the main signal.

4.3 Power Loss

When a pulse is generated by the AWG of a certain power, the final output of the board will be less powerful due to losses. Overall losses at different points during the process for a 12 MHz sinusoidal signal at 0.600 Vpp (Volt peak to peak) can be characterized and understood using measurements displayed in Figures 11 and 12.
The initial input power on the decibel scale can be calculated from the given value of x (in Vpp) by

$$\frac{\mathrm{Power}(\frac{x}{\mathrm{Vpp}})}{\mathrm{dBm}} = 10 \cdot \log(\frac{(\frac{x}{2})^2 \cdot 1000}{50\,\Omega}). \tag{3}$$

The wave is generated with a power of 2.5 dBm. It enters the IQ-mixer with a power of roughly -7.5 dBm, as in the setup for this measurement a 10 dB attenuator instead of 3 dB was installed in front of the I and Q channel. The calibration for the measurements in Figure 11 and 12 was adjusted to minimize the left IQ-sideband. Therefore the frequency to look at is the right sideband. As LO1 is set to 2 GHz at 15 dBm, the band comes out at a frequency of 2.012 GHz with a measured power of -10.5 dBm. This lies perfectly within the specified conversion loss[3] of typically 5.5 dB. An attenuation of 1 dB for the output of the switch comes from cables and elements between those two measured points. With LO2 set to 20 GHz at 18 dBm, the second up conversion done in the M8-mixer eventually converts the whole signal further up. One should note, that less than 18 dBm from the LO signal actually reach the mixer, due to non-negligible attenuation of 2-3 dBm for this measurement. This has in general to be accounted for when setting the LO's. As for every usual mixing process, the signal will be split into two sidebands after mixing in the M8-mixer. In the further discussion the focus will be on the right M8-sideband. Right after this second up conversion, the signal of interest has a frequency of 22.012 GHz and its power is measured to be -24 dBm. This constitutes a loss of 12.5 dB. Out of this, 6.5 dB can be amounted to the typical conversion loss of the M8-mixer[4]. As now the signal is in the high microwave regime, the other 6 dB are attenuated in the cable connecting the board and the spectrum analyzer. The actual signal should therefore be around -18 dBm. Eventually, a bandpass filter[5] and cables lower the power by another 4 dBm. So the total measured power of the frequency component which was aimed at is -28 dBm. Sub tracting 6 dB for the attenuation in the measurement cable yields a raw output power of approximately -24 dBm. Comparing the power of incoming I and Q signal with this particular band gives a total loss of roughly 21.5 dB. It is important to stress the fact, that this particular result only holds for the configuration of this measurement. Especially if the frequency of LO2 is changed, the losses will deviate from this value as can be seen in the next section. Nonetheless, the actual point to be made is, that all losses are well understood.
The 'maximum theoretical output power' of the targeted frequency shall denote the power, that can be acquired when raising I and Q input power to the 1 dB compression point. With these powers lying 13 dB below this point for the measurements mentioned above, the maximum output power for a 22 GHz up conversion of a 12 MHz signal is -11 dBm, which was also tested. Hence, attention has to be paid with the subsequent use of both 20 dB amplifiers.

[3]see appendix IQ-1545
[4]see appendix M8-0326
[5]see appendix FB-2400

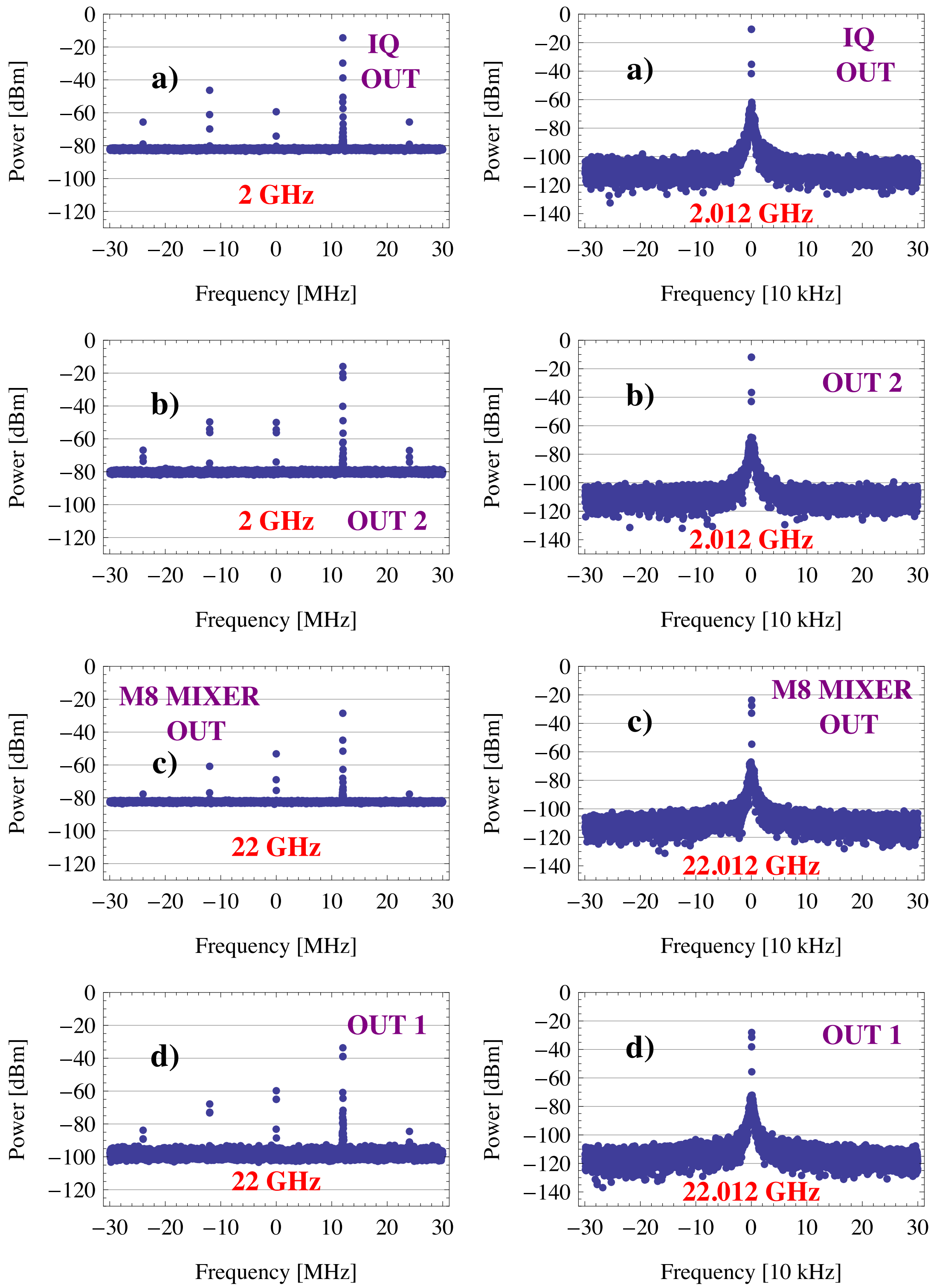

Figure 11: Spectrum at different stages of the up conversion of a 12 MHz sinusoidal wave. The left column displays a broader view centered around the expected output frequency. On the righthand side the power of the aimed at frequency can be tracked. For this measurement the right IQ-sideband and right M8-sideband was chosen. The LO's were set to 2 and 20 GHz respectively. a) Shows the spectra after IQ-mixer, b) after Out2, c) after M8 mixer, d) after Out1.

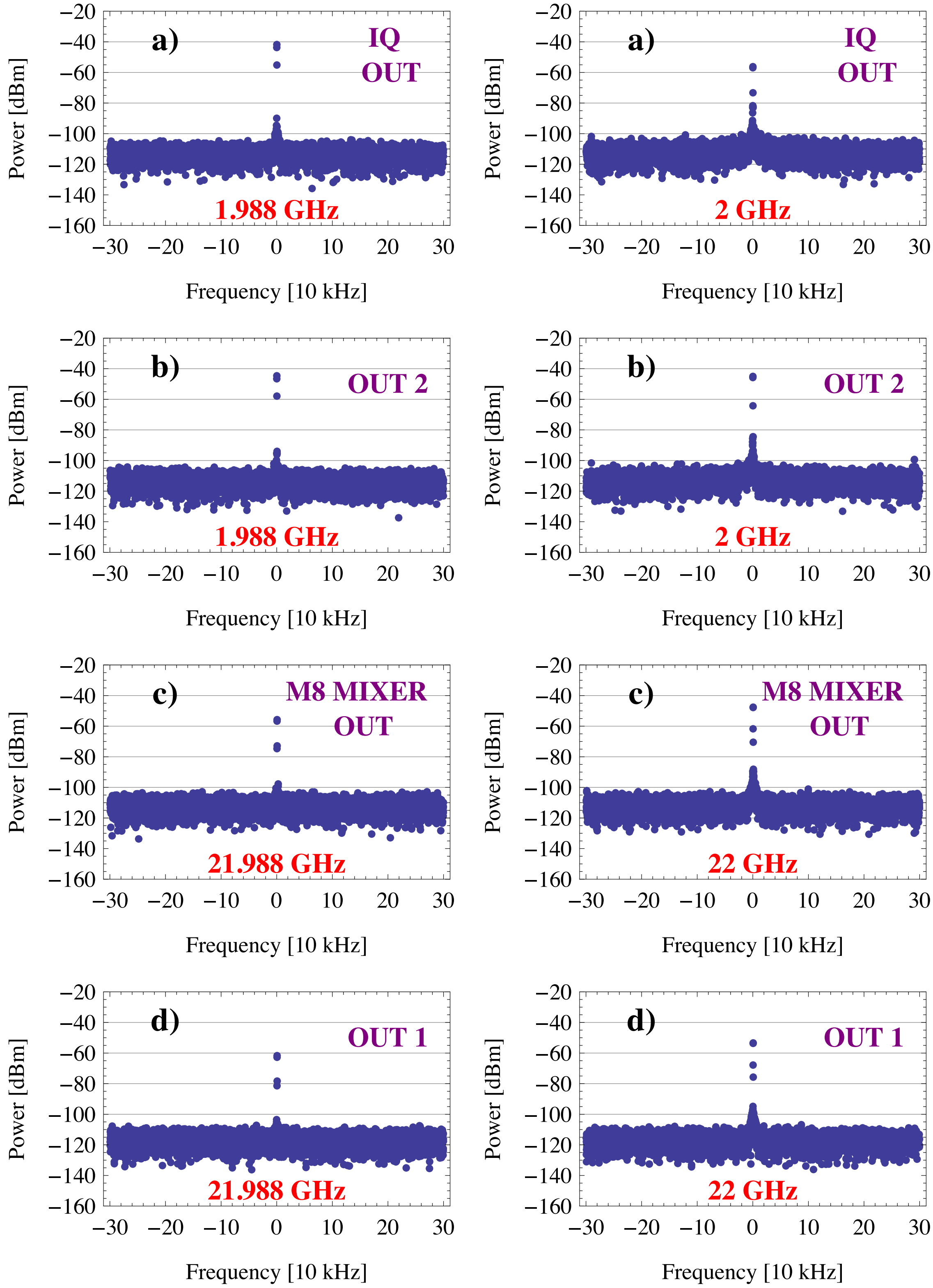

Figure 12: Development of unwanted IQ-sideband in the left column and leakage from LO1 on the right at different stages during the up conversion. Same settings as for Figure 11. No particularly good calibration was chosen. a) Shows the spectra after IQ-mixer, b) after Out2, c) after M8 mixer, d) after Out1.

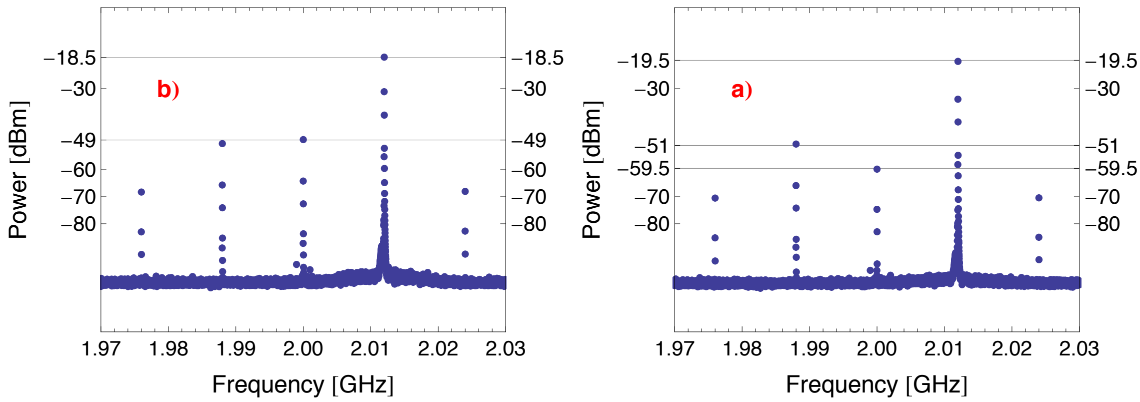

Figure 13: Power spectrum after switch b) without any filter installed between switch and IQ-mixer. a) with a 3 dB attenuator installed between switch and IQ-mixer. Up conversion settings as in Figure 11.

A subtle phenomenon arises when having a closer look at the left column of Figure 11. Somehow the proportion between LO1 peak and the left sideband is changed when going from the IQ-mixer through the switch to the output. It seems like the LO1 leakage peak actually gains power when going through the switch, which eventually makes it bigger than the left sideband. Possible explanations could be the existence of a standing wave between both elements, as well as some power leakage from the DC voltage applied to the switch. Figure 13 highly suggests, that this phenomenon vanishes, when an attenuating element is placed between IQ-mixer and switch. In the final configuration a filter at exactly this stage solves this problem.

4.4 Up converting pulses

The whole setup can be tested by up converting different pulse patterns. For this purpose the QUDEV 'cleansweep' measuring method, combined with an FPGA, is ideal. One starts out with a set of generated pulse patterns. Usually from each pattern to the next some parameter is changed. The Pattern is given to the AWG and a complete up conversion process takes place. The outcoming signal enters a down conversion board to down convert the signal to the 25 MHz regime. Amplitude and phase of the final signal are recorded for a certain period of time larger than the pulse duration. Using an FPGA with settings chosen, such that measured frequency and set frequency in the pulse pattern coincide. Before saving the data, digital filters can be applied. This helps avoiding an internal 'DC peak' phenomenon, which is sometimes necessary to filter away. Afterwards, the whole process is repeated with the next pulse pattern in the set. Thus, the result can be summarized in a three dimensional plot.
In Figure 14 the amplitude of gaussian pulses is plotted vs a line varying parameter on the y axis. Plots a), b) and d) of Figure 14 show the measured amplitude of pulses with varying width, phase and amplitude respectively. On the left side are the experimental results, whereas on the right side theoretical plots generated from the used wave patterns is shown. In order to exactly fit the measured pulsewidth, the theoreretical plots had to be stretched by 8%. It is not clear where this comes from, but it is for all three pulseshapes the same. Therefore it can only be a systematic problem. A Fourier transform of the measured data when the center frequency of a gaussian pulse is varied is shown in Figure 15 a). With ν being the frequency of the pulse pattern, the final signal in this measurement will internally arrive at the FPGA with a frequency of $\nu_R \equiv 25\,\text{MHz}+\nu$ for the right (R) and $\nu_L \equiv 25\,\text{MHz}-\nu$ for the left (L) IQ-sideband. As this is a real valued signal, the Fourier components c_f for each sideband contain an equal amount of $c_{\pm\nu_i}$. What then internally happens is a digital down conversion by 25 MHz of a complexified signal. This basically shifts down the Fourier coefficients for the digital signal according to $c_{R,L} = c_{\pm\nu_{R,L}-25\,MHz}$. As a result, the Fourier components for (R) can be found at (ν,$-50\,\text{MHz}-\nu$) and for (L) at ($-\nu$,$-50\,\text{MHz}+\nu$). As the FPGA has a measuring rate of 100 MHz, the maximum detectable frequency is ±50 MHz. Therefore, the -(50 MHz+ν) component of (R) will be folded back to 50 MHz-ν because of the Nyquist theorem. This perfectly fits the observation seen in Figure 15 a). It can be seen, that the IQ-mixer was calibrated to maximize the right sideband as indicated by the arrow, since the calibration was initially found with a 12 MHz pattern. This is again confirmed by the data, the left sideband being minimal at 12 MHz. Another aspect that can be seen from this image is that the calibration is not perfectly stable when varying the pulse frequency. This suggests to always find new calibration parameters when changing the frequency too much in the pulse pattern.

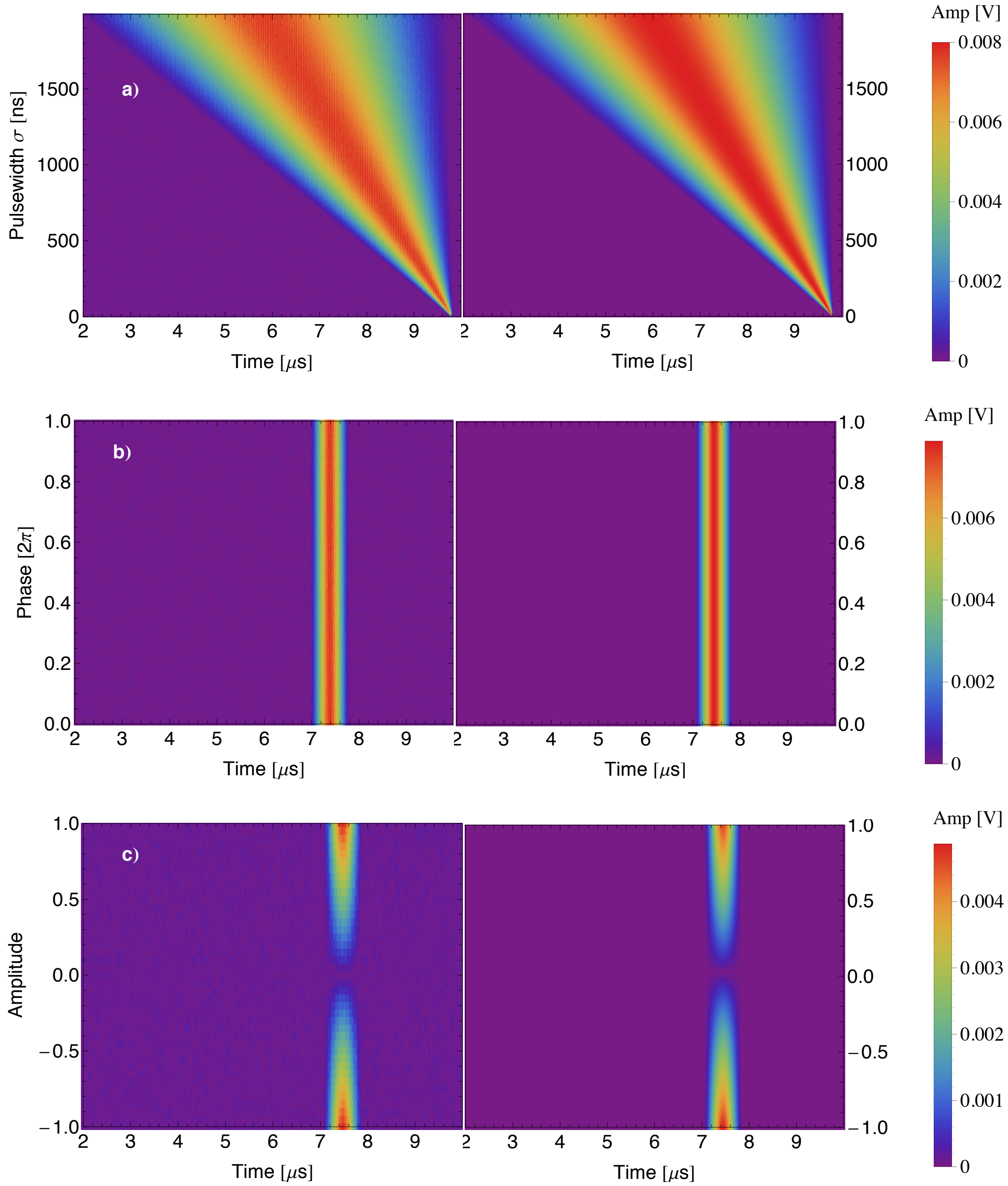

Figure 14: 22 GHz up converted pulses measured with an FPGA and the QUDEV 'clean-sweep' measuring program shown on the left. The intensity is plotted versus the swept parameters and time. On the right is a theoretical plot generated from the used wavepattern. Theoretical pulse needed artificial time duration stretch of 8% in order to fit the experimental results. a) 10 MHz gauss pulse with width $\sigma = 200$ ns, cheb(0.3,2,13)&4pt square digital filters. b) 10 MHz gauss pulse, cheb(0.3,2,13)&4pt square digital filter. c) Gauss pulse of width $\sigma = 200$ ns, no digital filter. d) 12 MHz gauss pulse, width $\sigma = 200$ ns, cheb(0.3,2,13)&4pt square digital filters.

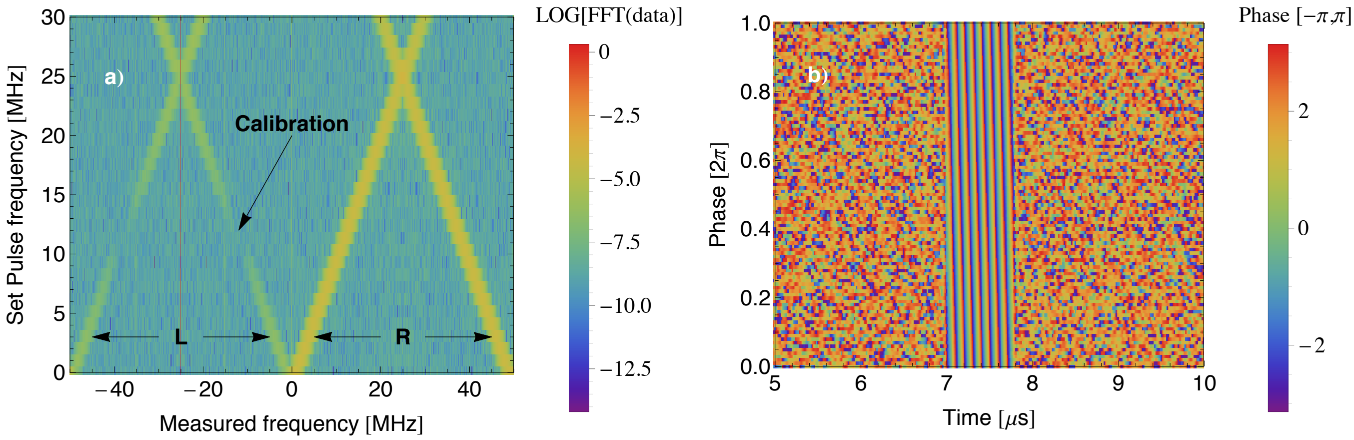

Figure 15: a) Logarithm of fourier transform of a gaussian pulse with varying center frequency. The calculated fourier frequency of the data is plotted on the x-axis versus the theoretical value on the y-axis. b) Phase of a phase varying pulse plotted over time versus the theoretical phase from the wave pattern.

In part b) of Figure 15 the phase variation of the phase varying pulse pattern is shown. During the time of the pulse a phase oscillation on the time axis is clearly visible. This is the change of phase due to time evolution. The changed parameter in the pattern, namely the starting phase , is hardly visible in this image. But when plotting the slice along the $t = 7.21\,\mu$s axis, this property can clearly be identified. Figure 16 confirms the correct mapping of the phase shift from input pattern to output signal.

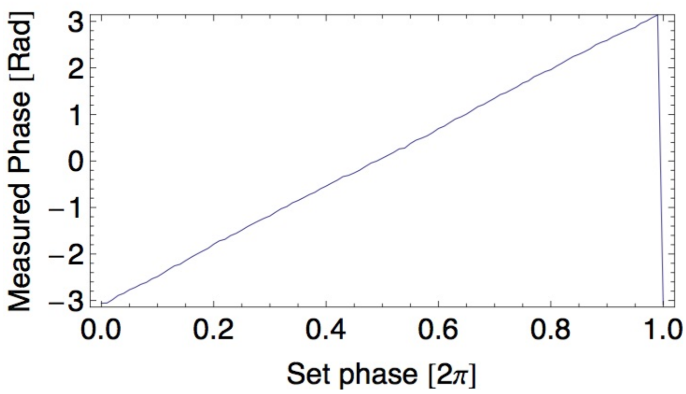

Figure 16: Gauss pulse from Figure 15 b) sliced along $t = 7.21\,\mu$s. Changing phase in pulse pattern equally shifts the measured phase.

An interesting aspect to investigate is the conversion efficiency. This is the property of the whole system consisting of up and down conversion board, cables etc. to efficiently transmit the initial pulse through the conversion processes. In order to get a rough idea of this characteristic, the amplitude of a 12 MHz pulse is measured for different up conversion frequencies η in Figure 17 a). A clear peak at 20 GHz marks a very good efficiency for this configuration. At both ends, the signal drops significantly into the noise. A better view on the development of this property along a slice at $t = 7.04\,\mu s$ is given in Figure 17 b).

The width of a square pulse is varied in Figure **??** a). A well known property of such

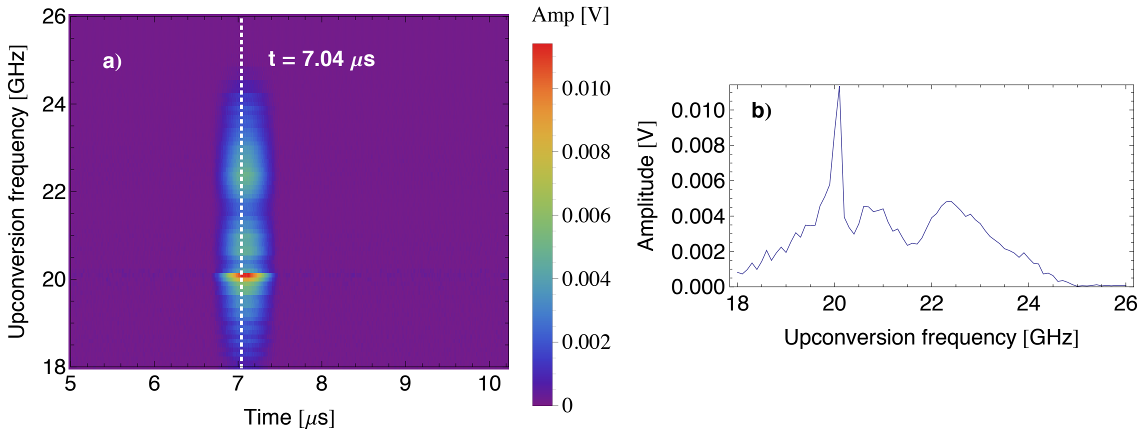

Figure 17: a) Amplitude of a 12 MHz gauss pulse, width $\sigma = 200\,\text{ns}$, for different up conversion frequencies η. b) Amplitude of such a Gauss pulse along the slice $t = 7.04\,\mu s$.

a pulse is the sinc function behavior of its Fourier transform. When taking the Fourier transform of the $\sigma = 0.4\,\mu$s slice, an actual sinc function can be observed and compared to theory. The pulse pattern for this particular slice can be used to calculate the theoretical Fourier transform of the initial pulse. Both, measurement and theory are displayed in Figure 18 b). The theoretical plot was scaled in order to fit the experimental results. A large 0 MHz contribution is measured in the actual signal. This comes from the fact, that the square pulse was DC pulse and thus the LO1 leak and the pulse leave the up conversion board at the same center frequency. When taking out the measurement point at 0 MHz ('DC peak'), the theoretical values can be scaled to fit the experimental results very well.
Overall it can be stated, that the output pulses are faithfully transmitted with respect to the generating pattern.

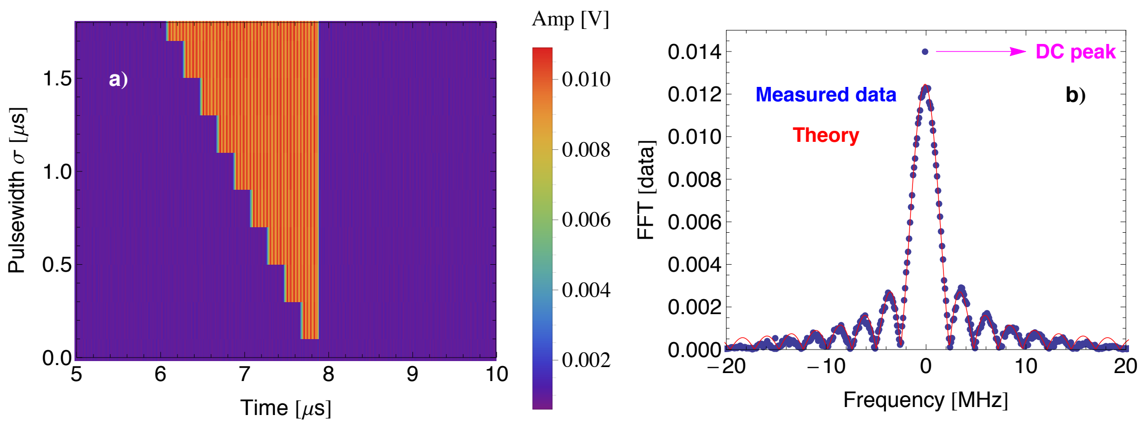

Figure 18: a) Logarithm of intensity along slice of image 3158 for $t = 7\,\mu$s. b) 'Measured data' denotes the intensity along the slice of image 3145 for $\sigma = 0.4\,\mu$s. The 'theory' data are generated from the fourier transform of the original wave pattern.

5 Conclusion

It was demonstrated, that the 'up conversion board' is a suitable device to generate arbitrary pulses at microwave frequencies between 3-26.5 GHz without difficulty. Furthermore, the engineering allows a clean spectrum, due to calibration and appropriate filters. The calibration can easily be done by using the intentionally provided second output port 'Out2'. As an example, the feasibility of a 60 dB contrast in the vicinity, as well as an overall 30 dB contrast for a 12 MHz signal, up converted by 22 GHz, was shown. The power loss from AWG to output signal depends on the up conversion settings, but can be explained by the effect of various components in the board. For the same signal as mentioned above, a power loss of approximately 21.5 dB is normal and an output power of -11 dBm can theoretically be achieved. Different pulse shapes are faithfully transmitted to the output. Varying different parameters in the pulse shapes are correctly mapped to the output, as can be seen from several examples. An instruction sheet with all relevant data, settings and 'howto' is provided in the appendix.

5.1 Outlook

The machinery for applying arbitrary pulses to the experimental setup now is ready to be used. It opens the door to various interesting measurements such as 'rabi oscillations', 'state tomography' and 'spin locking'. This could allow an even better understanding of the interaction between Rydberg atoms and their coupling to DC and AC electrical fields.

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A Components

1. IQ-1545/M from marki
2. M8-0326/NZ from marki
3. FB-2400 from marki
4. SLP-2400+ from minicircuits
5. TTL switch

ETH

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