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A quadrature frequency converter in a feedback loop of high
frequency cavities in the Proton Synchrotron at CERN.

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Życiorys

Urodziłem się 19 października 1989 r. w Warszawie. Cały proces edukacji przeszedłem w Warszawie, począwszy od Szkoły Podstawowej nr 76 im. 13 Dywizji Piechoty Strzelców Kresowych przez Gimnazjum nr 101 i IX Liceum Ogólnokształcące im. Klementyny Hoffmanowej. W 2008 roku rozpocząłem studia na Wydziale Elektroniki i Technik Informacyjnych Politechniki Warszawskiej. W 2010 roku odbyłem praktyki w firmie Pyrylandia, w roku 2011 byłem praktykantem w Wojskowym Instytucie Łączności, a od września 2011 roku do listopada 2012 roku byłem na stażu w Europejskim Laboratorium Fizyki Cząstek (CERN) pod Genewą, w Szwajcarii.

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Podpis

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Summary

This thesis presents the author's work during the internship at the European Laboratory for Particle Physics (CERN). The quadrature frequency converter is one of the modules that has been developed to upgrade the Proton Synchrotron RF system. Basic information about accelerators, fundamentals of IQ signal representation, mixing and phase shifting techniques are introduced. The development process of the converter is presented with the design details and measurements of the prototype board.

Kwadraturowy konwerter częstotliwości w pętli sprzężenia zwrotnego
rezonatorów Synchrotronu Protonowego w CERN.

Streszczenie

Niniejsza praca opisuje projekt autora zrealizowany podczas stażu w Europejskim Laboratorium Fizyki Cząstek (CERN). Kwadraturowy konwerter częstotliwości jest jednym z elementów opracowanych w ramach udoskonalenia systemu RF w Synchrotronie Protonowym. W pracy podane są podstawowe informacje o akceleratorach oraz podstawy reprezentacji sygnałów za pomocą składowych synfazowej i kwadraturowej. Omówione są także techniki mieszania sygnałów oraz przesuwania fazy. Opisany jest cały proces opracowania konwertera wraz ze szczegółami projektowymi i pomiarami wykonanego prototypu.

I am grateful to Dr. Heiko Damerau who supervised the internship at CERN for
introducing me to the world of accelerators.

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Chapter 1.

Introduction

One of the fundamental areas in physics is studying the structure of matter which consists of extremely small elements. The resolution of the measurement depends on the wavelength of the probe which has to be small compared to the size of the measured element [1]. The probe can be an electromagnetic radiation, but also a particles beam [2]. Because of the wave–corpuscle dualism we can assign the de Broglie wavelength to the particle given by:

$$\lambda_B = \frac{h}{p} = \frac{hc}{E} \quad \text{Eq. 1.1}$$

Increasing the particle energy shortens the assigned wavelength. As a result, the resolution of measurement is increased and we can observe smaller structures. Combining the de Broglie theory with the Heisenberg uncertainty principle brings the values of energy needed to explore the structure with a given resolution. Tab. 1.1 summarises required energies to explore given distances.

Δx (cm)	E	Tool
10^{-5}	2 eV	Microscopes
10^{-8}	2 keV	X rays
10^{-11}	$2 \text{ MeV} \simeq 4m_e$	γ rays
10^{-14}	$2 \text{ GeV} \simeq 2m_p$	Accelerators
10^{-16}	$200 \text{ GeV} \simeq 2m_{W,Z}$	Accelerators
10^{-17}	2 TeV	Accelerators

Tab. 1.1 Orders of magnitude of the minimum energy needed and tools used to explore distances of the order of Δx [3].

Observing the finest structures is not the only aspect of elementary particle physics. Scientists are able to create new particles. Energy of colliding primary particles can be converted to a mass of newly created one according to Einstein equation:

$$E = mc^2 \quad \text{Eq. 1.2}$$

Consequently, particles at high energy can be transformed to more massive new particles.

Because of these two reasons, high energy particles are required in physics. Natural sources of such particles (e.g. cosmic rays) give low intensive and uncontrollable flux. To

perform experiments, high energies are achieved in machines called accelerators. The most famous accelerator is the LHC (Large Hadron Collider), operated at CERN (European Organization for Nuclear Research). LHC does not operate alone, but is a part of the CERN accelerator complex shown in Fig. 1.1

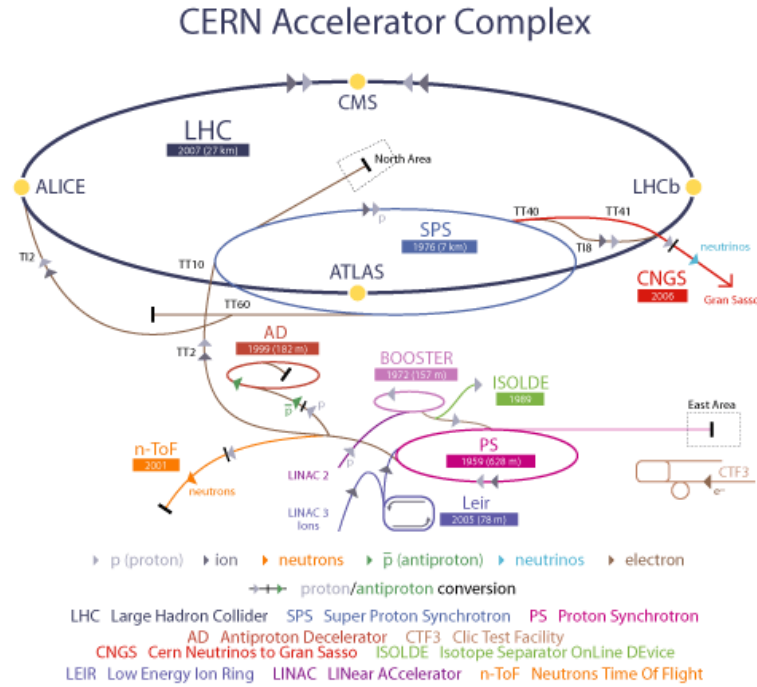


Fig. 1.1 CERN accelerator complex [4].

There are two basic types of accelerators: linear and circular. Particles pass through the linear accelerators only once, while they make many turns in the circular ones. In general, circular accelerators achieve high energies efficiently, as the particles pass through the accelerating stations multiple times. Particles are accelerated by longitudinal electric fields, while the direction of travel is determined by the magnetic field. The Proton Synchrotron (PS) at CERN is a type of circular accelerator where the accelerating voltage and the bending magnetic field are synchronised with the particles revolution frequency. As a result, the accelerating electric field is alternating. Voltage is applied to the particles in radio frequency resonators, called cavities.

Increasing the energy and intensity of particle beams in accelerators is crucial for physics experiments, but different phenomena limit the performance. One of the problems is beam loading. *Beam loading usually refers to the effects induced by the passage of the beam in the radio frequency cavities* [5]. *The beam current induces a voltage in the cavity which may become even larger than the one induced by the generator* [6]. In the main accelerating cavities in the PS this voltage may easily reach even 8 kV in amplitude [7] with

tens of kW of RF power induced by the beam. There are several methods to overcome the problem: cavity detuning, feedforward and feedback. As the long delays might be unavoidable a direct wideband feedback system would not damp the voltage sufficiently due to the limitation of the gain. The feedback bandwidth would be too narrow to damp the beam induced voltage in the cavity at harmonics of the revolution frequency. However, a special feedback system has been developed which takes advantage of the fact that a delay of one revolution period is acceptable in the feedback loop. It is called a one-turn-delay feedback. A gap return signal from the cavity is sent to a comb filter with gain maxima at harmonics of the revolution frequency and applied to the cavity with a delay equal to the revolution period. The phase can be adjusted at each revolution harmonic such that it cancels the beam induced voltage.

An existing feedback board for the one-turn delay feedback of the 10 MHz cavities in the Proton Synchrotron operates in base-band, but it cannot be directly applied to 40 MHz and 80 MHz cavities. To simplify the design and production process, one-turn delay feedback modules for the higher frequency cavities use the frequency conversion technique. The induced signal in cavities is down converted by an IQ demodulator to the in-phase and quadrature signals at intermediate frequency (e.g. 10 MHz), then applied to the one-turn delay feedback module. The feedback signal is then up converted back by an IQ modulator to the desired frequency and applied to the cavity.

This thesis describes the author's work on the quadrature frequency converter during the internship at CERN from September 2011 to November 2012. It starts with the introduction explaining the reasons for developing the project. The subsequent chapter provides information about IQ signal representation and the influence of IQ mismatch on the performance of the one-turn-delay feedback module. Required parameters of the frequency converter are estimated. In the third chapter the basic mixing and quadrature phase shift generating techniques are described. These techniques are fundamental in the process of quadrature frequency conversion. After finding the parameters and considering available techniques, the overview of the components market is given in the fifth chapter. The sixth chapter describes the prototype board and discusses the measurements. The summary is given in the seventh chapter. As there are many ECL components used in the design, a short description of the technology and its termination techniques can be found in Appendix II.

Chapter 2.

RF signals in accelerators

This chapter describes basic parameters and signals that can be found in RF systems of the Proton Synchrotron. Moreover, a spectrum of the beam induced voltage in the cavity is explained.

2.1 Beam current spectrum

The beam of particles is accelerated in the presence of a longitudinal electric field. This is done by applying a voltage across a gap in the vacuum pipe. The voltage has to be alternating, otherwise there would be no overall acceleration: before passing the gap, the particles are accelerated, behind the gap they are decelerated. As the Proton Synchrotron is a circular accelerator, the particles pass through the same gap many times. Care must be taken to ensure, that the particles always see an accelerating voltage across the gap. Therefore, the frequency of the gap voltage have to be equal to the revolution frequency or to its integer multiple. RF systems are often described with a **harmonic number** [8]:

$$h = \frac{f_{RF}}{f_{rev}} \quad \text{Eq. 2.1}$$

Where:

f_{RF} is the RF frequency;

f_{rev} is the revolution frequency.

To maximise the accelerating voltage at the gap, it is surrounded by a resonator (cavity) tuned to the given RF frequency.

The accelerated beam of charged particles is an electric current. At a given place, the beam current consists of pulses delivered at each passage. Such a signal is periodic with the revolution frequency. Due to the spread in momentums of particles, the pulses are not Dirac deltas, but they have a finite width [9]. Such a pulse is called a **bunch**.

Many functions can be used to describe the bunch. A good order of approximation, especially in estimating the performance of the frequency converter, is achieved with

a Gaussian model. Parameters of the bunch are length (σ or its multiple) and amplitude. A bunch signal used to determine the beam induced voltage in the cavity is shown in Fig. 2.1

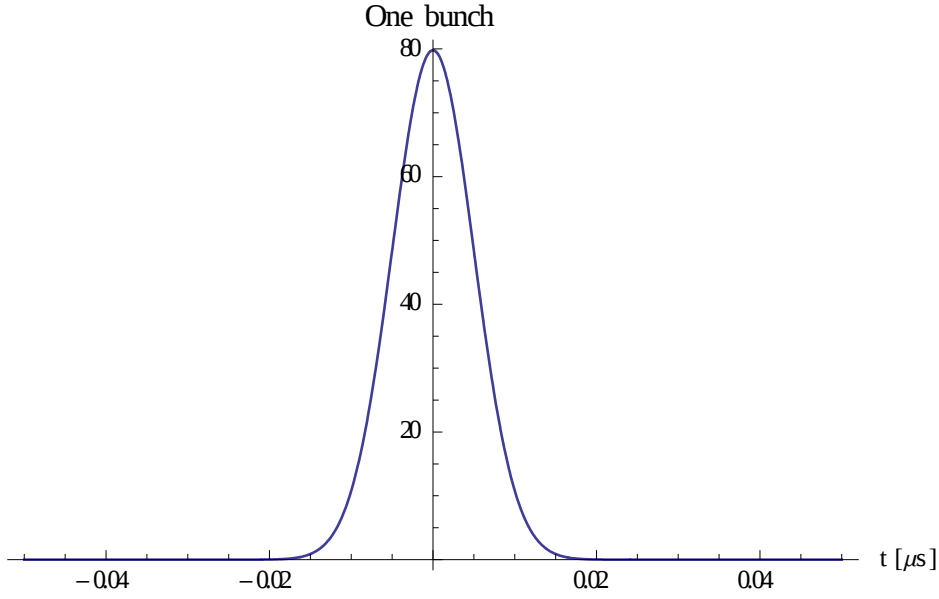


Fig. 2.1 One bunch, with a 4σ length equal to 20 ns.

Bunches are contained in **RF buckets** [8]. The length of the bucket is determined by the RF frequency (a harmonic number). Usually not all the created buckets are filled with bunches. Such a situation is shown in Fig. 2.2

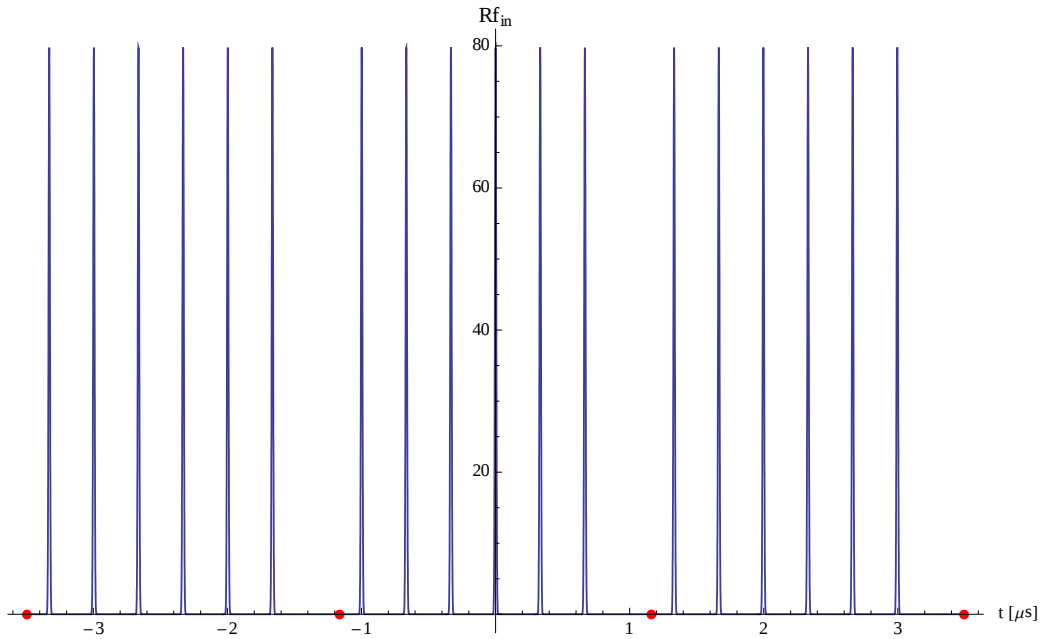


Fig. 2.2 A beam current signal consisting of 7 buckets, 6 bunches. The revolution frequency f_{rev} is 476.82 kHz, the harmonic number h is 7, which means that f_{RF} is 3337.74 kHz.

The beam current spectrum of the signal from Fig. 2.2 is shown in Fig. 2.3 and Fig. 2.4. The spectrum consists of peaks at harmonics of the revolution frequency. Every 7th peak is

stronger. As can be seen in Fig. 2.4, the spectrum lowers with some roll-off. The wider (the longer) the bunch is, the narrower the spectrum is – roll-off is higher. If the pulses were Dirac deltas, the spectrum would consist of peaks of the same amplitude.

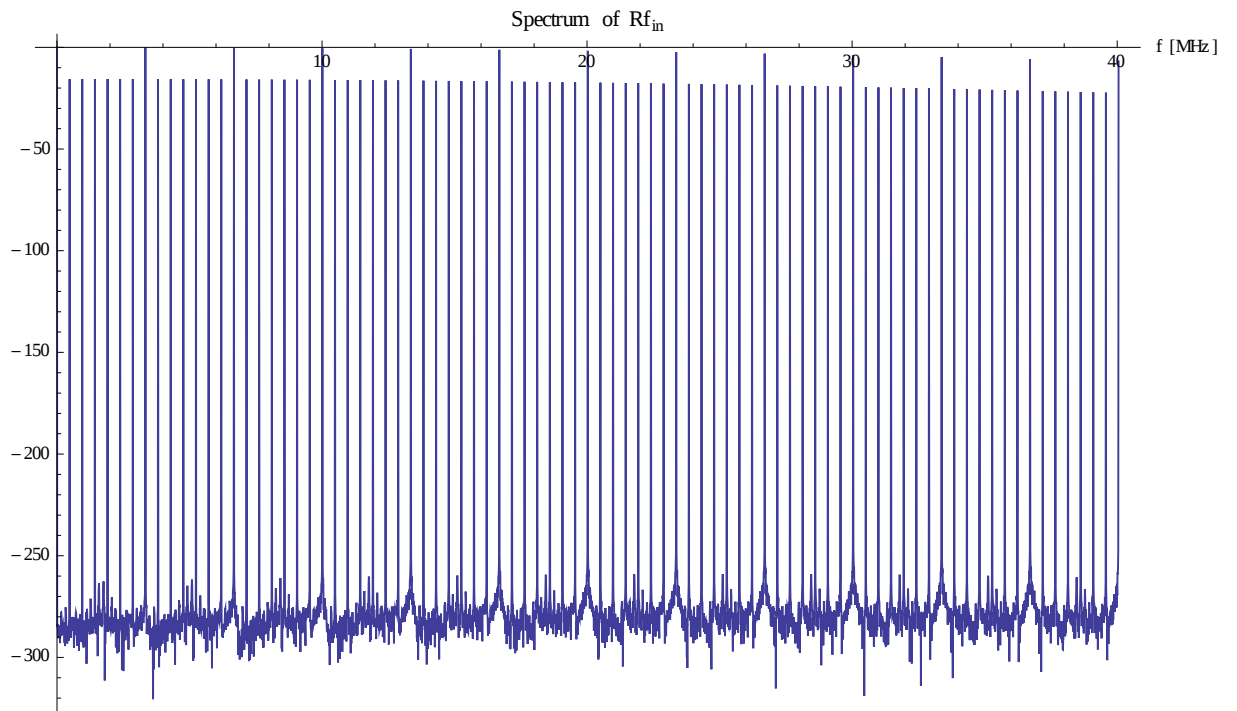


Fig. 2.3 A beam current spectrum shown in a narrow frequency range.

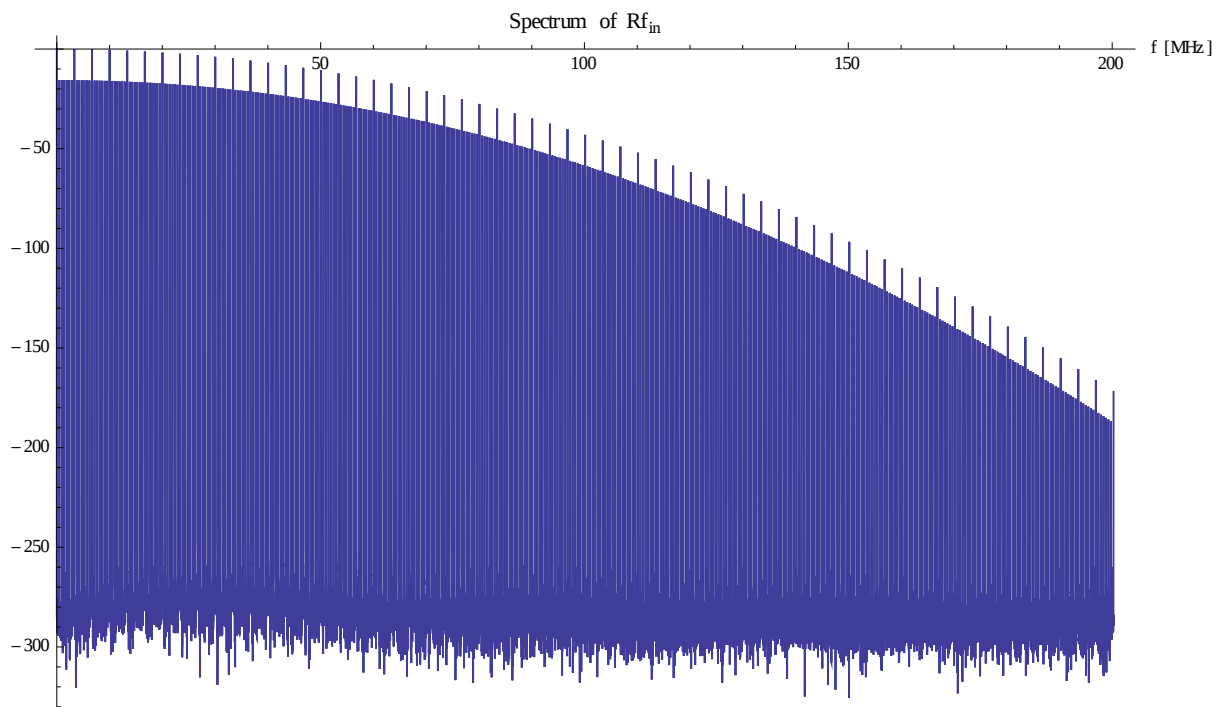


Fig. 2.4 A beam current spectrum shown in a wide frequency range.

2.2 Cavity transfer function

For frequencies in the neighbourhood of the fundamental resonance an RF cavity can be described by an equivalent circuit consisting of an inductance L_2 , a capacitor C and a shunt impedance R_S [6] as shown in Fig. 2.5.

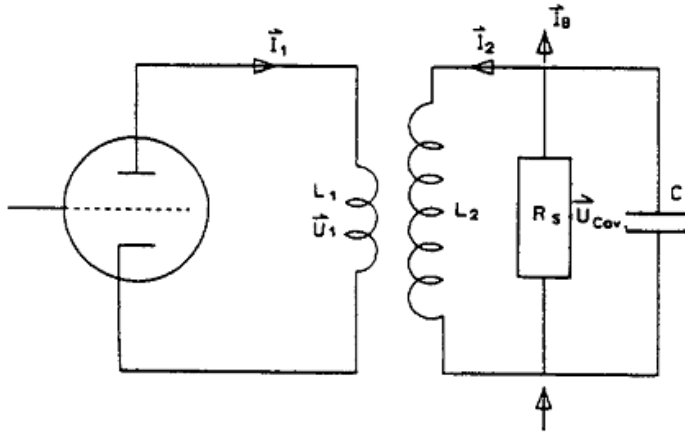


Fig. 2.5 Equivalent circuit of a resonant cavity near its fundamental resonance [6].

This is a standard parallel RLC resonant circuit, which can be also characterised with the shunt impedance R_S , quality factor Q and the resonant frequency f_0 . For the purposes of designing the frequency converter, the real cavity transfer function was approximated with this simple model. The parameters of the Proton Synchrotron 40 MHz cavity are:

- $R_S = 600 \text{ k}\Omega$;
- $Q = 18200$;
- $f_0 = 40.055 \text{ MHz}$.

The resulting transfer function is shown below.

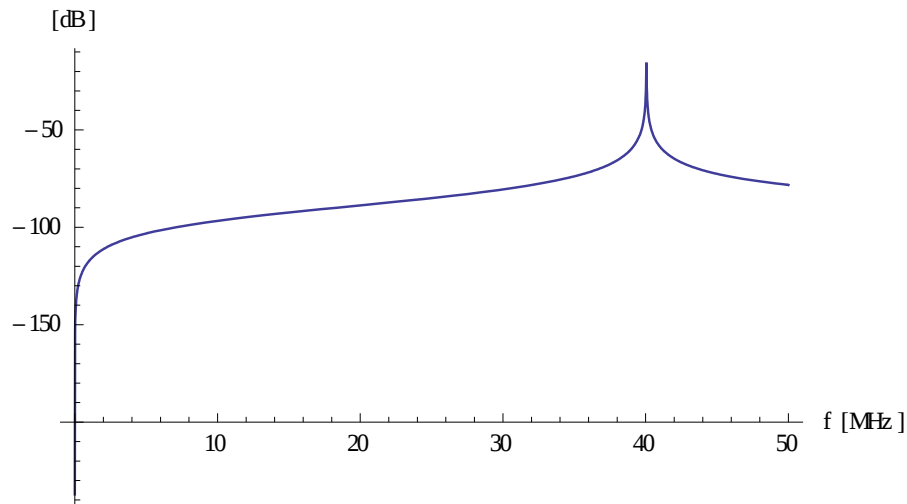


Fig. 2.6 Transfer function of the 40 MHz cavity.



Fig. 2.7 Cavity “Tjitske” SS77 in the Proton Synchrotron. The nominal f_0 is 40.055 MHz ($h = 84$).

2.3 Beam induced voltage in the cavity

The beam, passing through the cavity sees a high impedance. From the simple equation $U = I \cdot R$, the beam current induces a voltage in the cavity. Its spectrum is given by the product of the beam spectrum and the cavity transfer function [6]. The beam induced voltage in the cavity spectrum, based on previous examples, is shown in Fig. 2.8.

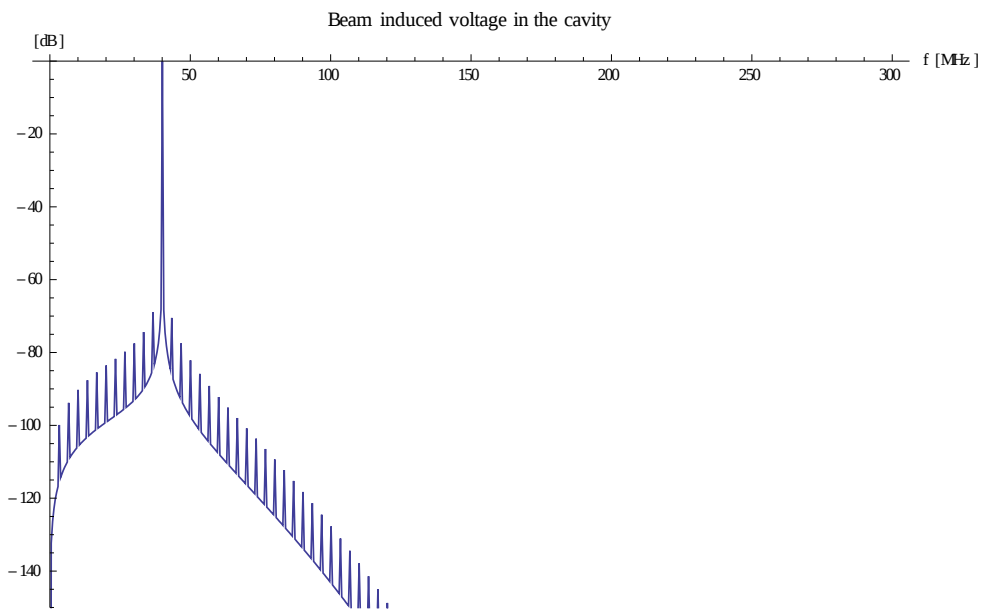


Fig. 2.8 Beam induced voltage in the cavity which is perfectly tuned to the 84th harmonic number.

Chapter 3.

IQ signal representation

Quadrature representation of signals is widely used in many systems. It can be found in communication systems because it offers a convenient way to modify the parameters of a carrier signal. Demand for increasing a capacity of radio channels require the use of simultaneous amplitude and phase modulation which is easily achievable with IQ modulators. Another advantage of IQ modulators is a possibility of creating image rejection mixers that lower the level of unwanted output signals in modulators and increase the performance of receivers. IQ signal processing is not only used in telecommunication systems, but also in accelerators RF systems. It is because of a useful feature that a signal amplitude and phase can be measured with respect to the reference signal [10].

3.1 IQ plane

Every sinusoidal signal with given amplitude, frequency and phase can be represented as a combination of two signals in quadrature [11]:

$$A \cos(2\pi ft + \varphi) = A \cos(2\pi ft) \cos(\varphi) - A \sin(2\pi ft) \sin(\varphi) \quad \text{Eq. 3.1}$$

$$I = A \cos(\varphi) \quad \text{Eq. 3.2}$$

$$Q = A \sin(\varphi) \quad \text{Eq. 3.3}$$

$$A \cos(2\pi ft + \varphi) = I \cos(2\pi ft) - Q \sin(2\pi ft) \quad \text{Eq. 3.4}$$

Where:

I is the amplitude of the in-phase component,

Q is the amplitude of the quadrature component.

From above equations one can easily derive formulas for a amplitude and phase of the signal on the basis of amplitudes of its I and Q signals (minding the signs):

$$A = \sqrt{I^2 + Q^2} \quad \text{Eq. 3.5}$$

$$\varphi = \arctan\left(\frac{Q}{I}\right) \quad \text{Eq. 3.6}$$

According to Eq. 3.5 and Eq. 3.6, an amplitude and instantaneous phase can be modified by changing amplitudes of IQ signals. Frequency is a derivative of phase, thus it can be also modified by IQ amplitudes. Amplitude modulation is much easier in realisation than phase modulation and this explains a popularity of IQ modulators.

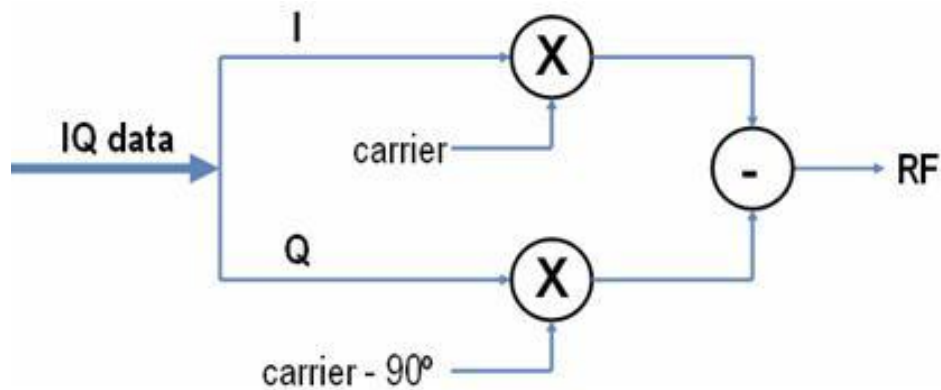


Fig. 3.1 Block diagram of an IQ modulator [11].

A state of a sine wave can be represented by a vector in a complex plane (Fig. 3.2). The length of the vector is the amplitude, an angle between the vector and the horizontal axis is the phase. Vector will be rotating around the origin with a frequency of the signal. If the phase measurement is done in reference to the sine wave of frequency equal to the one which is represented in a plane, the plot will be a stationary point instead of a rotating vector. IQ representation of a sine wave is a translation of vector parameters in a polar plane to a Cartesian coordinate system.

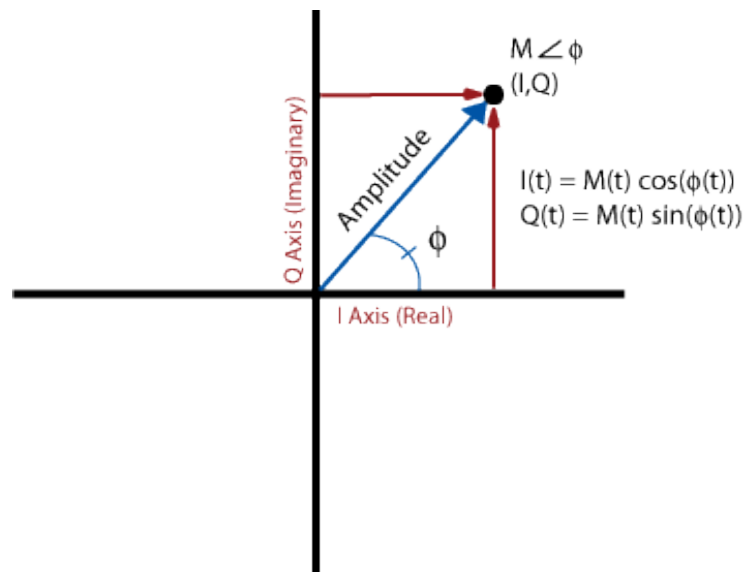


Fig. 3.2 Sine wave represented in a complex plane [11].

3.2 Errors in IQ modulation

3.2.1 Distorted signal spectrum

Quadrature frequency conversion requires symmetrical mixing of two signals. Imperfections in conversion paths (different parameters of transistors in mixers, not equal length of traces) may result in amplitude and phase imbalances of IQ signals. The frequency converter in the project firstly down converts the signal, afterwards up converts. What happens when a signal is not ideally modulated and demodulated? According to Fig. 3.3, after double frequency conversion, an output signal is given by Eq. 3.7.

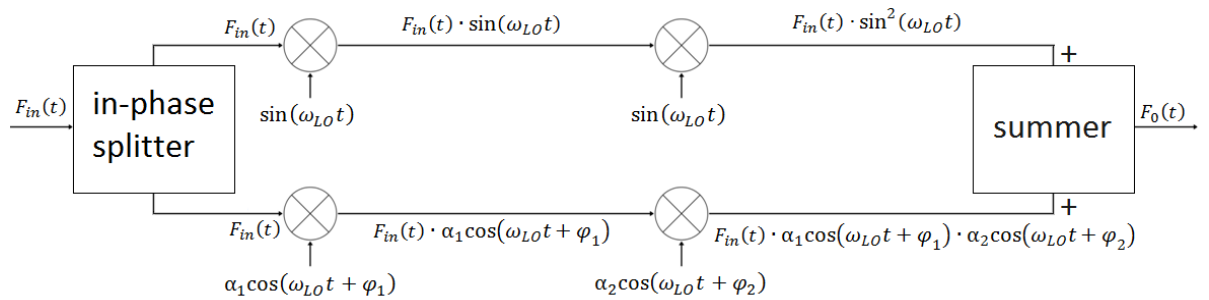


Fig. 3.3 Block diagram of double frequency conversion.

$$F_0(t) = [\sin^2(\omega_L t) + \alpha_1 \alpha_2 \cos(\omega_L t + \varphi_1) \cos(\omega_L t + \varphi_2)] F_{in}(t) \quad \text{Eq. 3.7}$$

Where:

$F_{in}(t)$ - input signal;

$F_0(t)$ - output signal;

ω_L - angular frequency of a heterodyne signal;

α_1, α_2 - amplitude mismatch introduced by 1st and 2nd frequency conversion;

φ_1, φ_2 - phase mismatch.

If there is no mismatch ($\alpha_1 = \alpha_2 = 1, \varphi_1 = \varphi_2 = 0$), the input signal is multiplied by $\sin^2(\omega_L t) + \cos^2(\omega_L t)$ which equals to 1, thus the signal is not modified.

If the mismatch is small, as shown in Appendix I, the output signal can be expressed as:

$$F_0(t) = \left\{ 1 + \frac{1}{2} [(\varepsilon_1 + \varepsilon_2) + (\varepsilon_1 + \varepsilon_2) \cos 2\omega_L t - (\varphi_1 + \varphi_2) \sin 2\omega_L t] \right\} F_{in}(t) \quad \text{Eq. 3.8}$$

All the terms in brackets, despite the first one, describe an error signal $F_E(t)$.

$$F_E(t) = \frac{1}{2} [(\varepsilon_1 + \varepsilon_2) + (\varepsilon_1 + \varepsilon_2) \cos 2\omega_L t - (\varphi_1 + \varphi_2) \sin 2\omega_L t] F_{in}(t) \quad \text{Eq. 3.9}$$

To find the spectrum, the error signal can be transformed to (Appendix I):

$$F_E(t) = \left\{ \frac{1}{2}(\varepsilon_1 + \varepsilon_2) + \frac{1}{4}[(\varepsilon_1 + \varepsilon_2) + j(\varphi_1 + \varphi_2)]e^{j2\omega_L t} + \frac{1}{4}[(\varepsilon_1 + \varepsilon_2) - j(\varphi_1 + \varphi_2)]e^{-j2\omega_L t} \right\} F_{in}(t) \quad \text{Eq. 3.10}$$

From properties of Fourier transform:

$$e^{2\pi j a x} f(t) \rightarrow \hat{f}(\xi - a) \quad \text{Eq. 3.11}$$

In our case it means that the unwanted spectral component is an original spectrum gained by a factor of $\frac{1}{2}(\varepsilon_1 + \varepsilon_2)$ and moved left and right at the double heterodyne frequency, with the same gain given by $\frac{1}{4}\sqrt{(\varepsilon_1 + \varepsilon_2)^2 + (\varphi_1 + \varphi_2)^2}$ but with a different phase. This can be shown on the charts made with Mathematica software.

Fig. 3.4 presents the ideal spectrum of induced voltage in a cavity (detuned), Fig. 3.5 - the spectrum of output signal from mismatched device. To facilitate the analysis, coloured circles have been added. Red ones correspond to positive frequency part of spectrum and blue ones to negative. In the second chart, four parts of spectrum are marked.

- 1 - the movement of the negative part of an ideal spectrum on the right, at the distance of double heterodyne frequency;
- 2 – as 1, but the positive part;
- 3 – movement of the negative part on the left;
- 4 – movement of the positive part on the left;

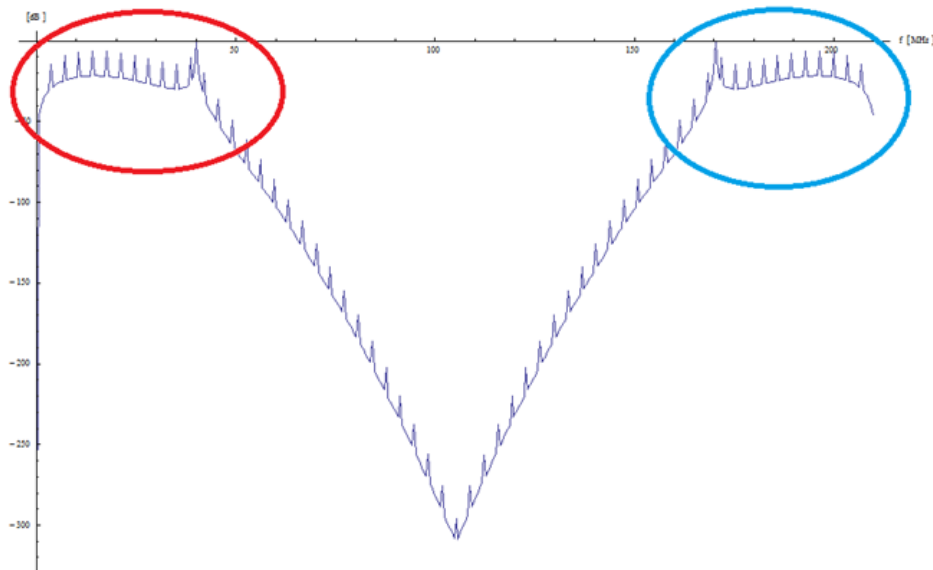


Fig. 3.4 Spectrum of an undistorted induced voltage in a cavity.

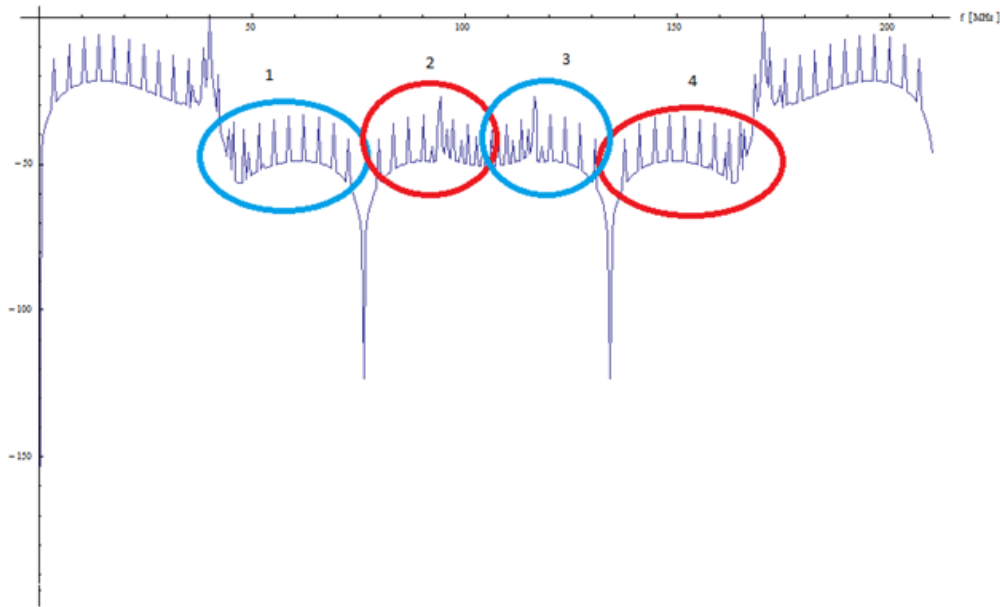


Fig. 3.5 Spectrum of an induced voltage in a cavity, distorted by IQ mismatch.

3.2.2 EVM – Error Vector Magnitude

Mismatch in phase and amplitude between I and Q signals distorts the signal. To characterise the influence of both types of errors with one characteristic measure, the error vector magnitude (EVM) was introduced. This measure is mainly used in digital radio communication systems but is useful in any system that uses quadrature signals. The error vector is defined as a vector in the IQ plane between the reference signal and measured (received) signal [12]. It is shown in Fig. 3.6.

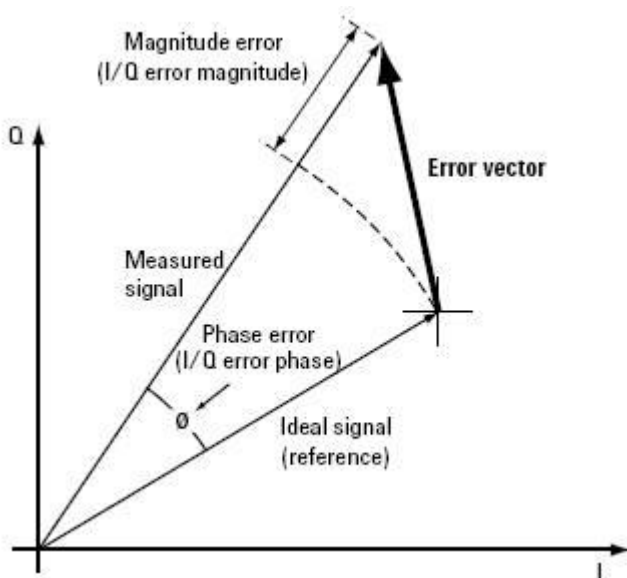


Fig. 3.6 Visualisation of EVM.

To fully characterise the performance of a quadrature system, one needs the information about both magnitude and phase errors. The knowledge of only one type of error is useless, as there can be a situation where e.g. magnitude error is negligible, but a phase error is so big that system does not work correctly. EVM is a useful measure, because it is sensitive to both types of errors. It gives a good overview of the system performance with a single parameter.

3.2.3 IQ mismatch and the performance of the one-turn-delay feedback

The beam current spectrum consists of peaks at the harmonics of the revolution frequency with little information between them [7]. The spectrum of induced voltage in cavities is a product of multiplying a beam spectrum by a frequency response of cavity. As a result, a significant part of the spectrum will be located at the revolution frequency harmonics close to the cavity resonant frequency. To simplify the task, the analysis of frequency converter chain in a one-turn-delay feedback module can be limited to this frequency range.

The influence of IQ mismatch on the beam induced voltage in cavities was analysed with a Mathematica notebook. The notebook allow a user to define all required parameters of beam such as pulse length, harmonic number, number of bunches and revolution frequency. Cavity is characterised by a shunt impedance, quality factor and resonant frequency. Real cavities do not resonate on a single frequency, but also have higher order modes. However, damping antennas are used to damp the modes [8; 13] and for IQ mismatch analysis a single resonance frequency model is fully sufficient. For a given amplitude and phase mismatch of demodulator and modulator, the notebook calculates the spectrum of the distorted signal and its differences from the undistorted spectrum. Distortion is measured by magnitude error, phase error and EVM. In addition to an error spectrum over a whole frequency range calculated for a single set of parameters, notebook generates plots of error measures for harmonics centred around the resonance of the cavity as a function of phase and amplitude mismatch. The user can also specify an intermediate frequency, which is defined as a harmonic number of the revolution frequency.

During the acceleration in the Proton Synchrotron the velocity of particles changes by 15 % [7]. Consequently, the revolution frequency increases. However, the cavity remains

tuned to the fixed frequency, thus the beam induced voltage changes the character as the revolution frequency increases. For a lower velocity, the spectrum around a resonant frequency consists of peaks at similar level. When particles travel faster, the cavity is tuned close to a revolution frequency harmonic and there is one strong peak in the spectrum. From the IQ mismatch point of view, the signal is more distorted when the cavity is detuned. For the tuned cavity, EVM takes significant values only at the resonant frequency.

Set of plots was prepared for two revolution frequencies: the lower 429.138 kHz and the nominal 476.82 kHz. The whole spectrum was calculated for following parameters:

Demodulator's amplitude error: 0.3 dB

Demodulator's phase error: 1°

Modulator's amplitude error: 0.5 dB

Modulator's phase error: 2°

Heterodyne frequency for a detuned cavity was 44.6304 MHz, for a tuned 49.5893 MHz. In both cases it is a 104 harmonic number.

3.2.3.1 Plots for a detuned cavity

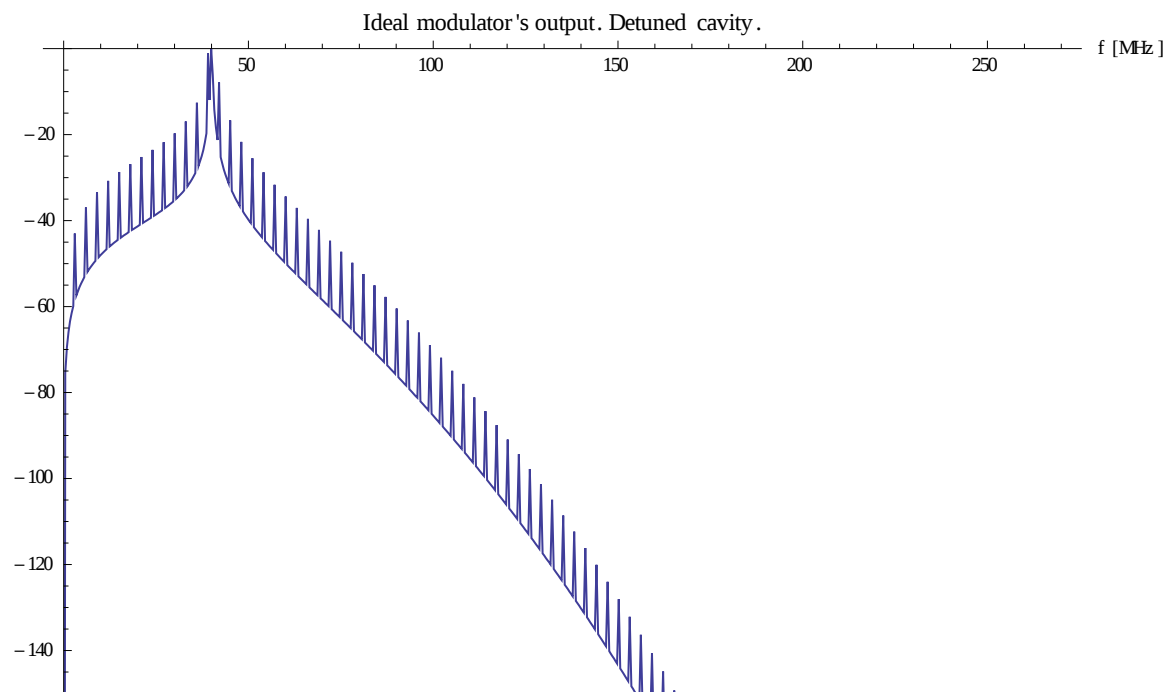


Fig. 3.7 Undistorted spectrum of beam induced voltage in a detuned cavity.

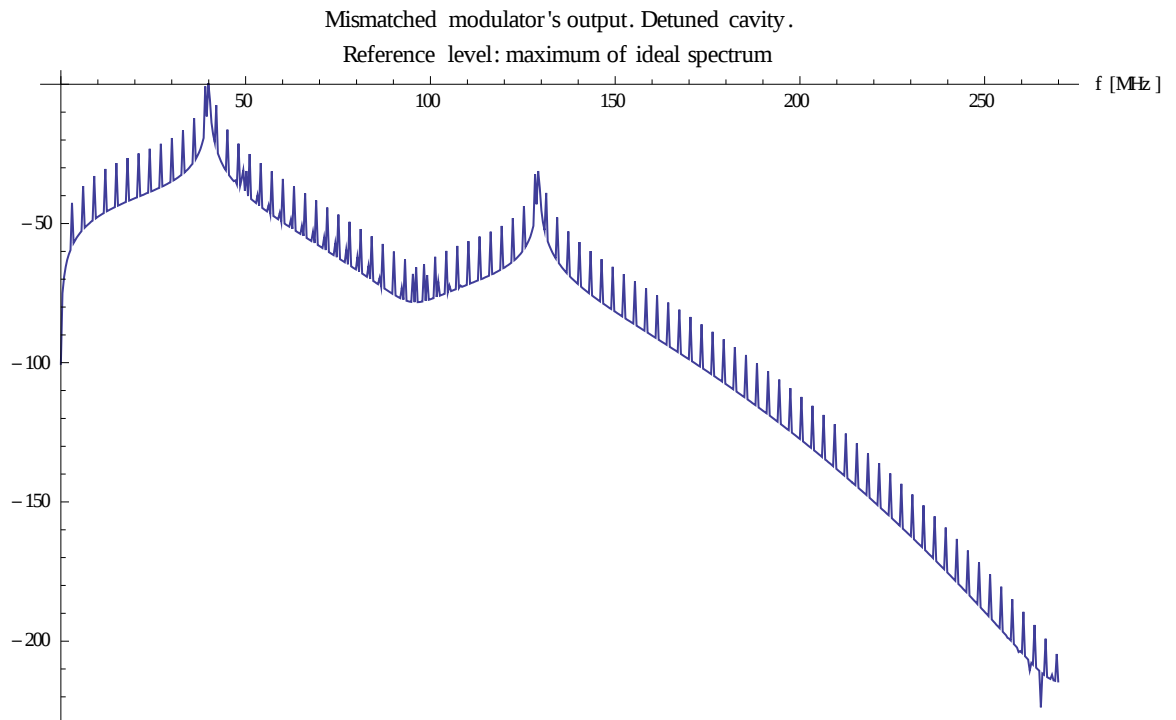


Fig. 3.8 Distorted spectrum of beam induced voltage in a detuned cavity.

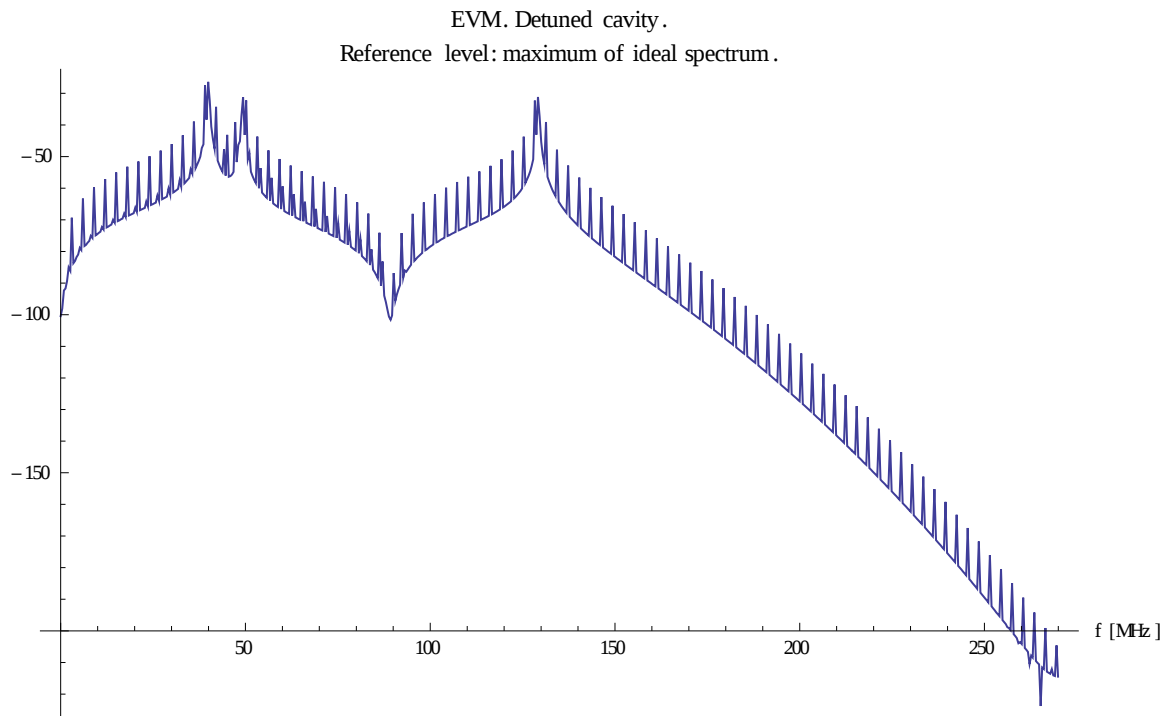


Fig. 3.9 EVM in a detuned cavity.

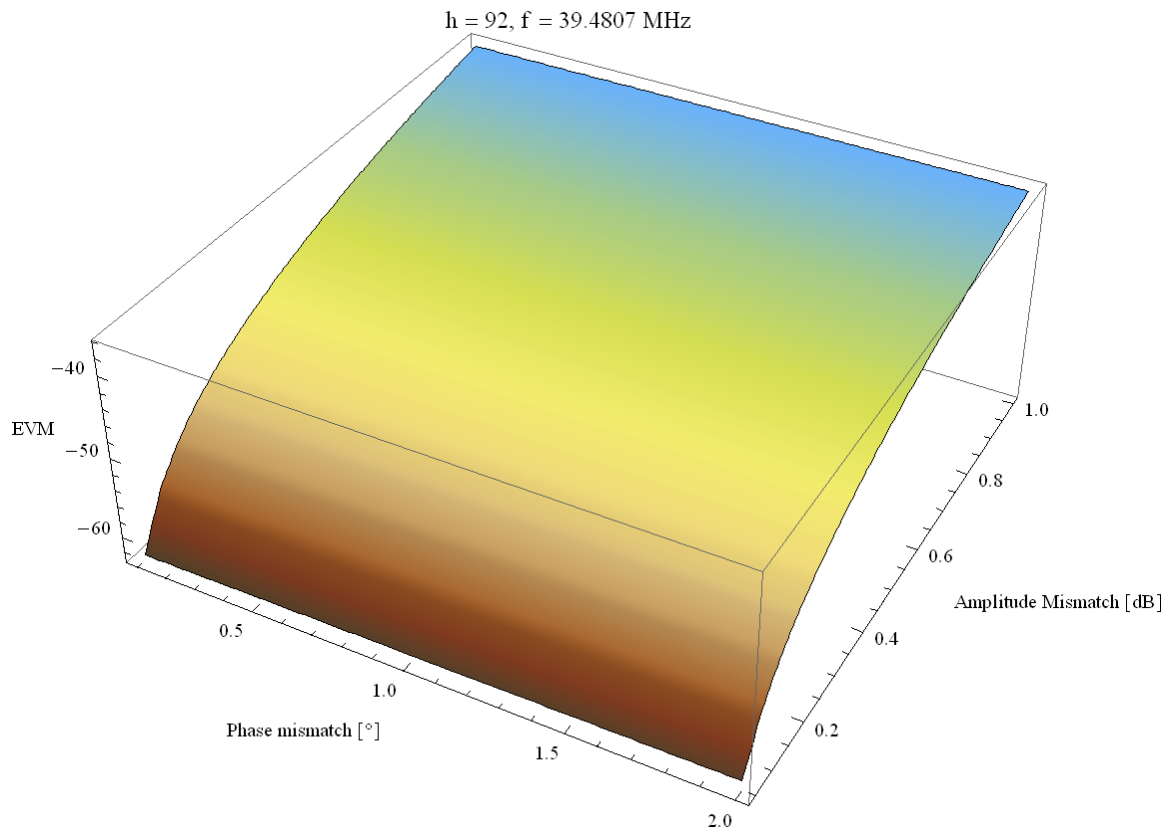


Fig. 3.10 EVM in a detuned cavity vs. phase and amplitude mismatch at $h = 92$.

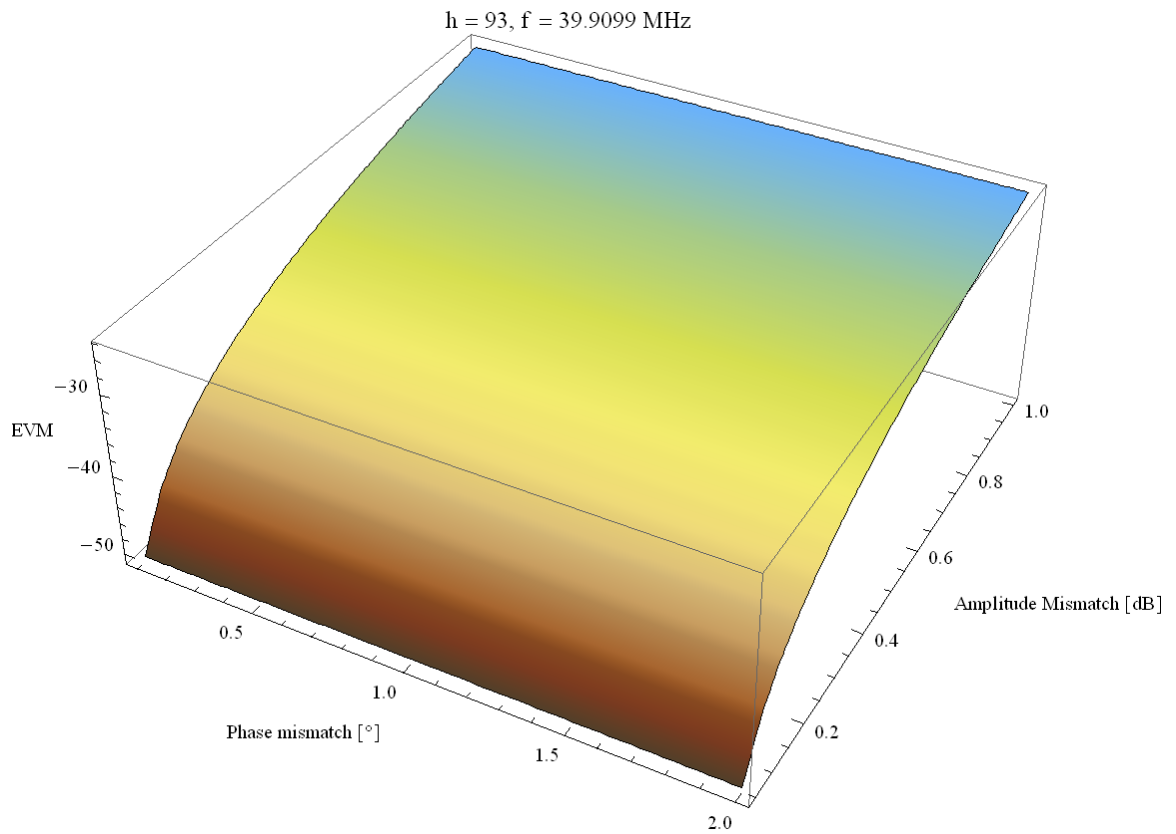


Fig. 3.11 EVM in a detuned cavity vs. phase and amplitude mismatch at $h = 93$.

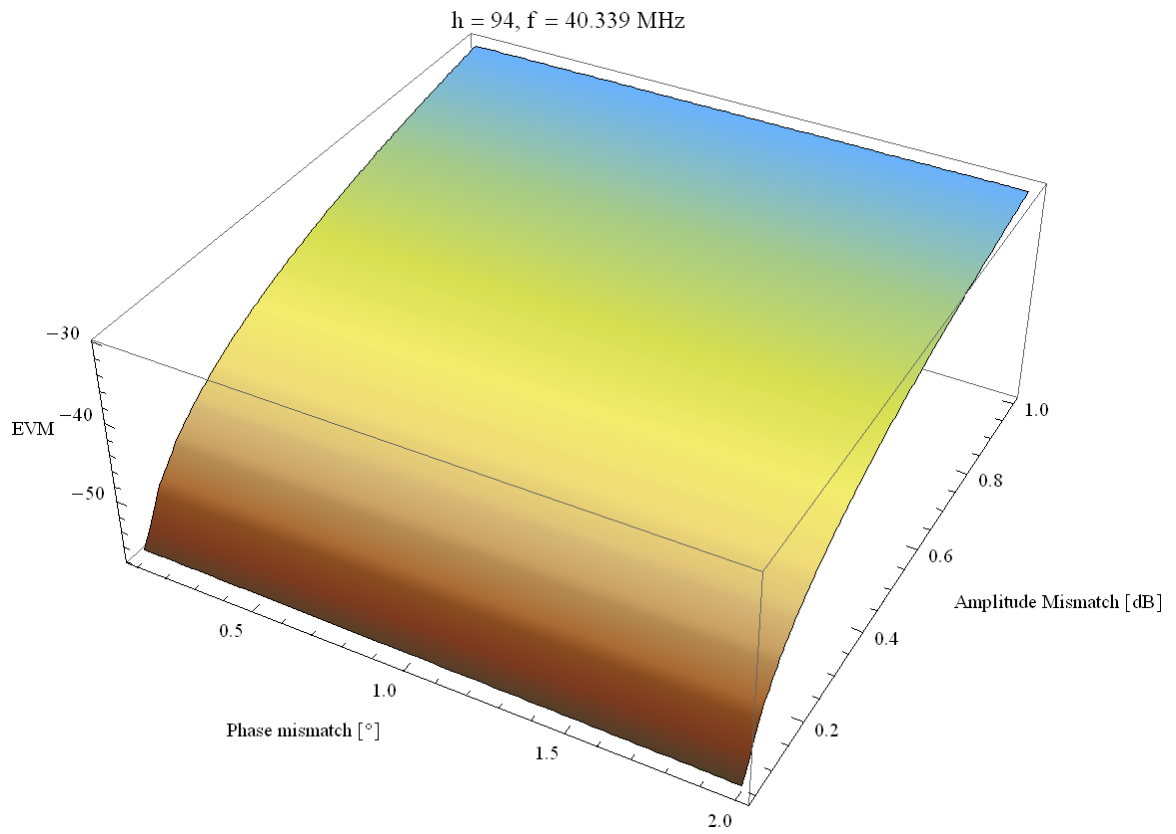


Fig. 3.12 EVM in a detuned cavity vs. phase and amplitude mismatch at $h = 94$.

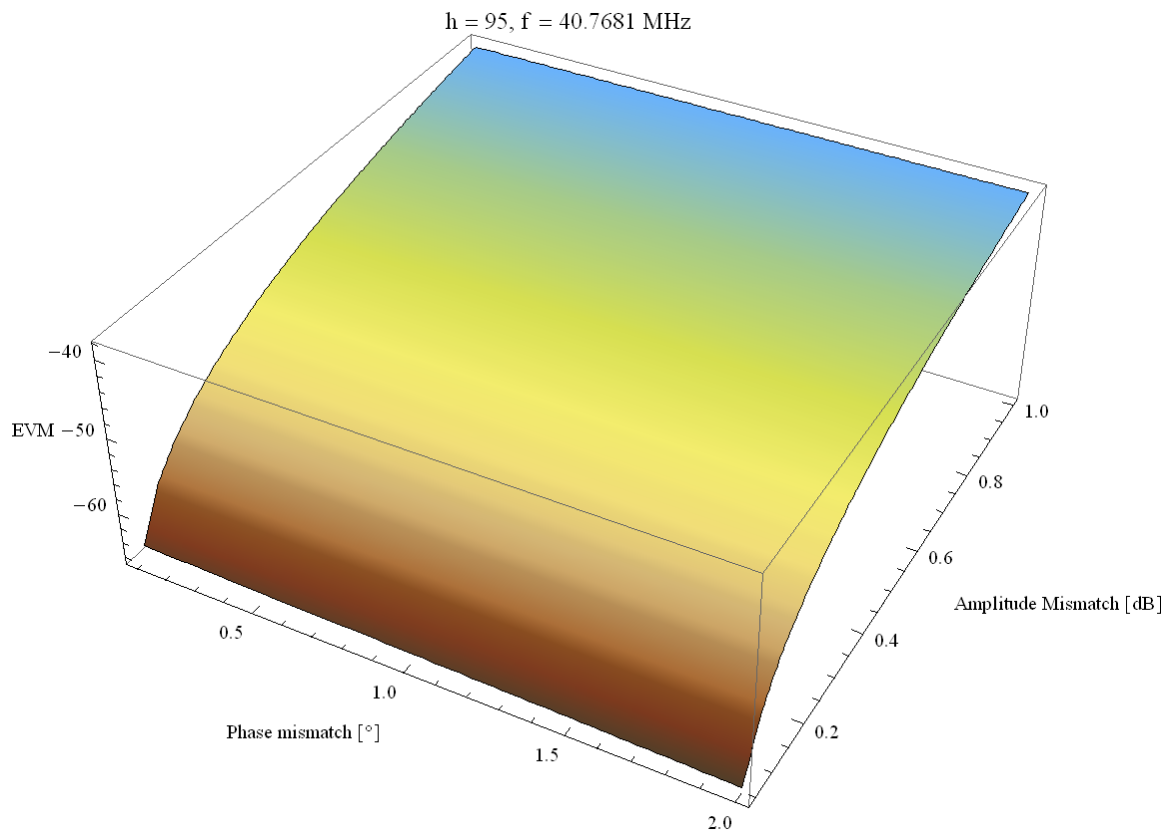


Fig. 3.13 EVM in a detuned cavity vs. phase and amplitude mismatch at $h = 95$.

3.2.3.2 Plots for a tuned cavity

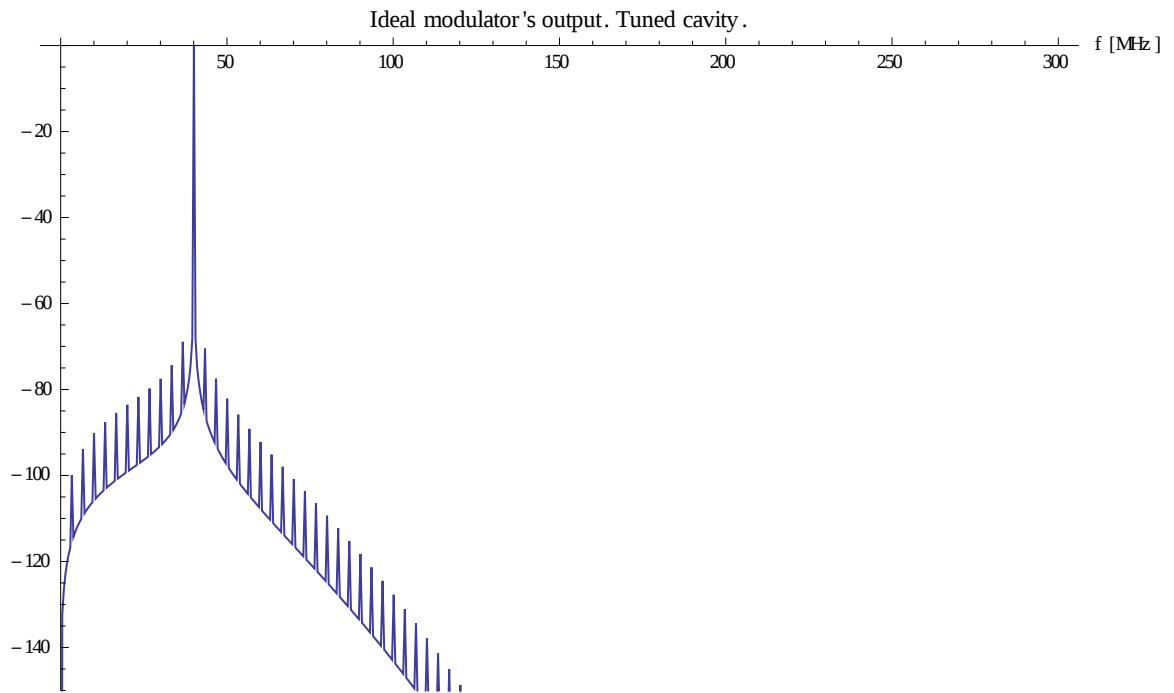


Fig. 3.14 Undistorted spectrum of beam induced voltage in a tuned cavity.

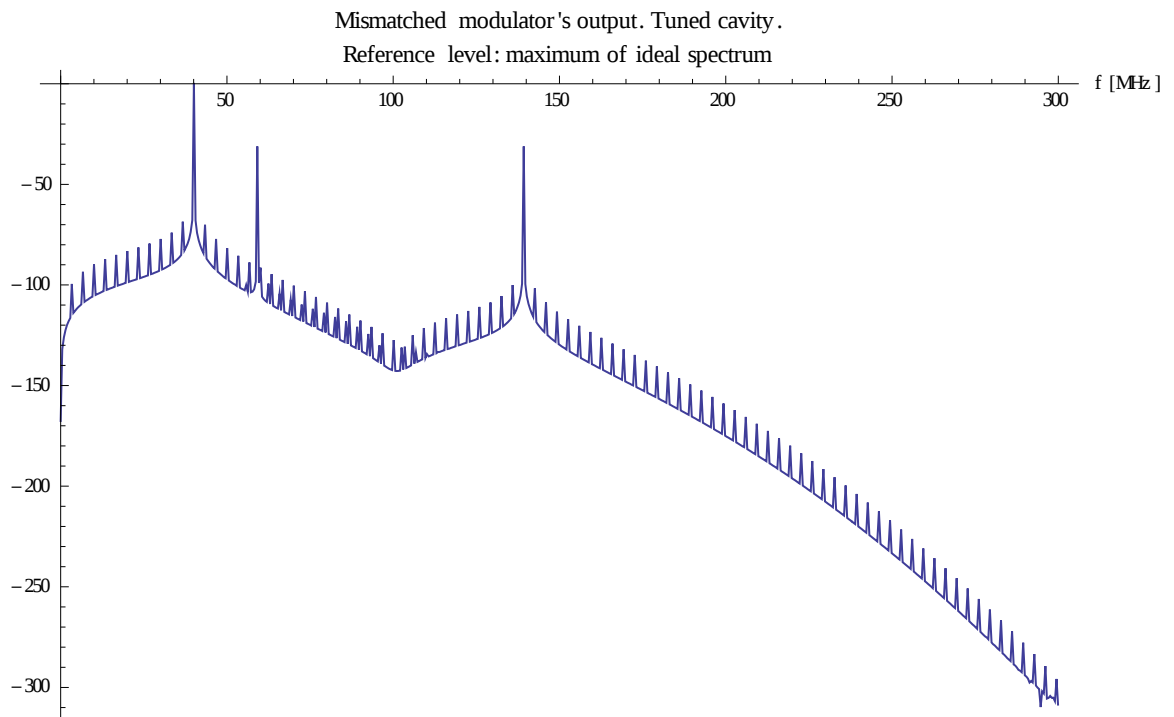


Fig. 3.15 Distorted spectrum of beam induced voltage in a tuned cavity.

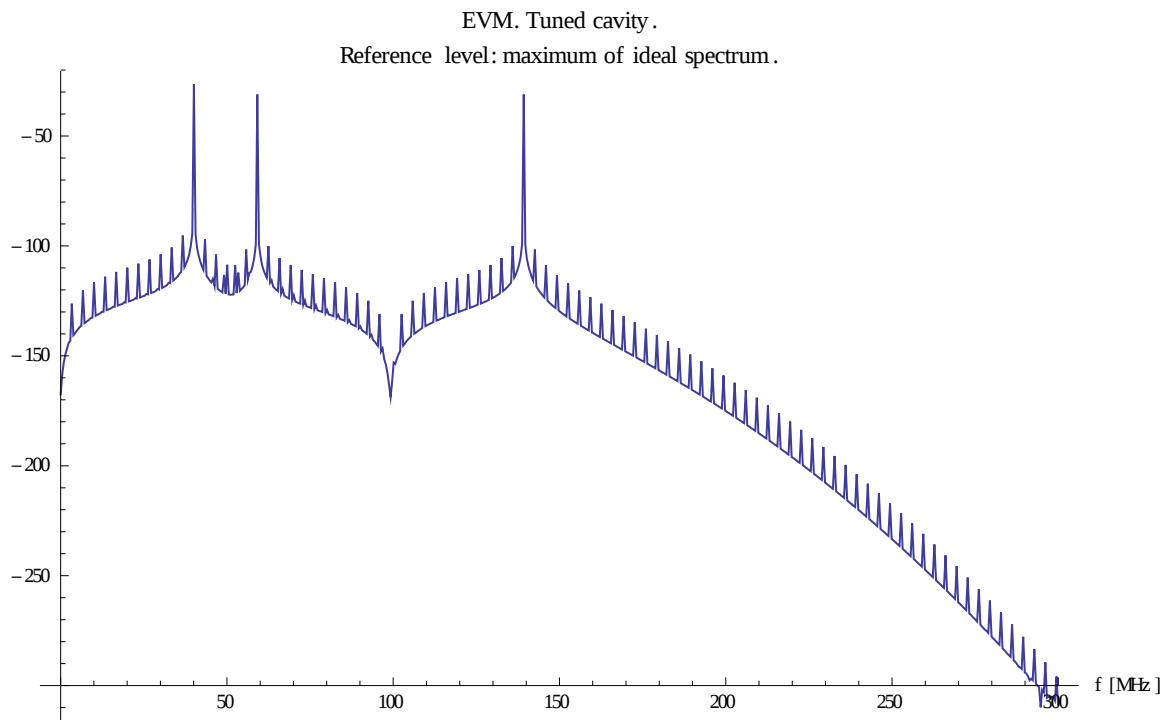


Fig. 3.16 EVM in a tuned cavity.

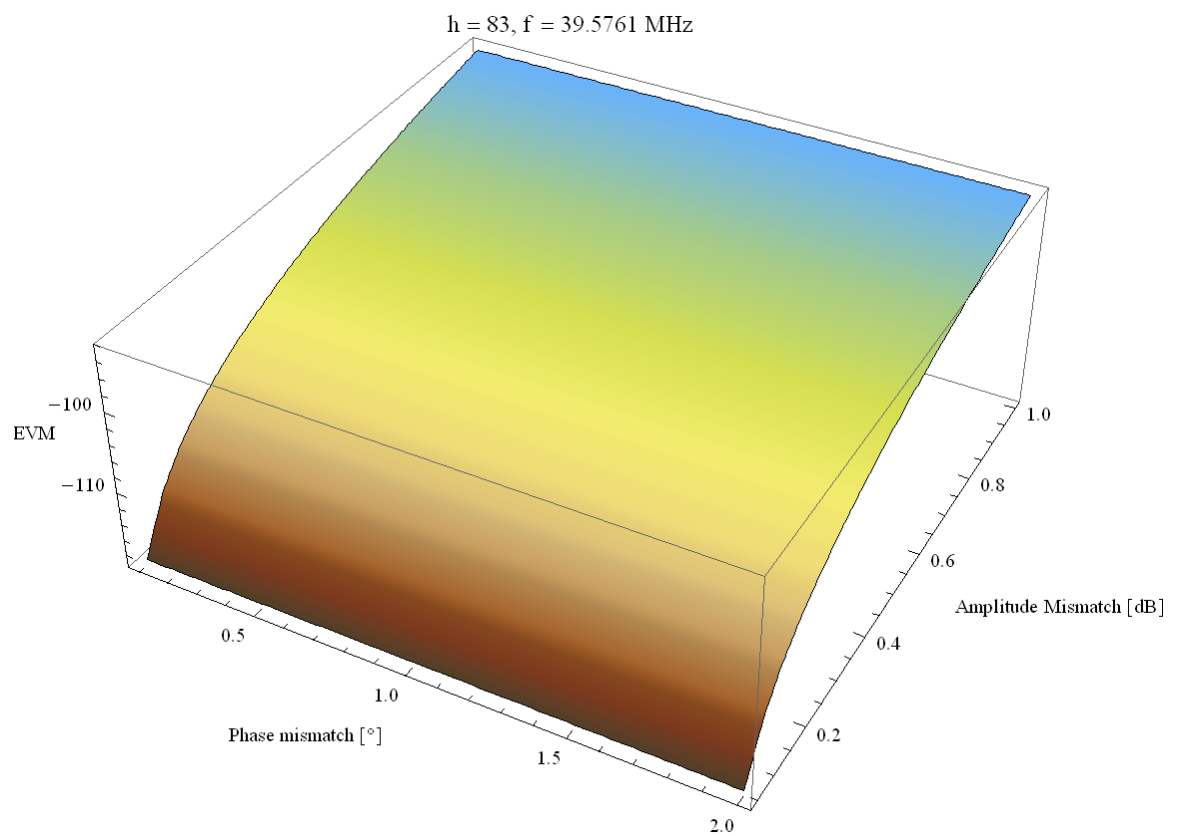


Fig. 3.17 EVM in a tuned cavity vs. phase and amplitude mismatch at $h = 83$.

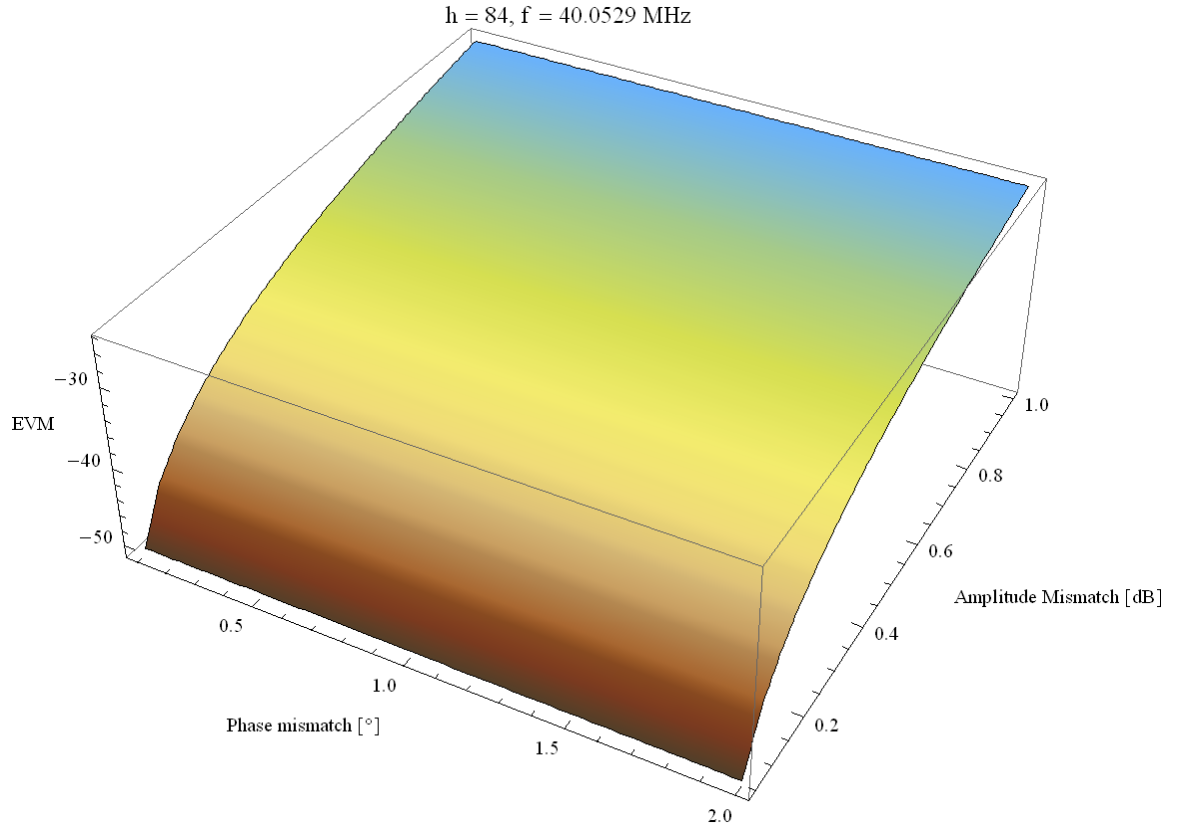


Fig. 3.18 EVM in a tuned cavity vs. phase and amplitude mismatch at $h = 84$.

3.2.4 Required parameters of the frequency converter

Fig. 3.7 and Fig. 3.14 show that spectrum of beam induced voltage in cavities consists of significant peaks around the resonant frequency. The figures also show that the system is more sensible to the mismatch when the cavity is detuned. Plots of EVM vs. phase and amplitude mismatch of IQ signals at harmonics around the resonant frequencies (Fig. 3.10 to Fig. 3.13, Fig. 3.17 and Fig. 3.18) are useful to estimate the maximum allowed mismatch in a frequency converter. Assuming that EVM on the level of -30 dB is acceptable, the frequency converter IQ phase mismatch should be smaller than 2.0° and amplitude mismatch should be less than 1.0 dB .

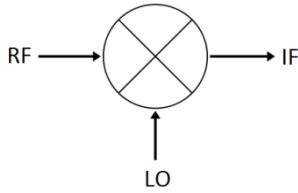
Chapter 4.

Basic techniques

Even though the one-turn-delay feedback module operates in digital domain, analog pre-processing has to be applied prior to analog-to-digital conversion due to limited ADCs performance [14]. One of the pre-processing functions is to down-convert signal frequency. As the feedback module operates on IQ signals it is convenient to translate the frequency with a quadrature frequency converter. Such a converter requires the use of mixers for frequency translation and a phase shifter to drive the mixers with LO signals in quadrature. Several analogue techniques for signal mixing (frequency shifting) and quadrature generation are discussed in this chapter.

4.1 Mixing techniques

Designers of linear devices, such as amplifiers, put effort into minimising level of intermodulation distortion. Nevertheless there are devices where the non-linearity is a principle of operation. In mixers, the intermodulation products are not distortion, but a useful signal. A non-linear device driven by two signals of different frequencies generate a signal at the sum and difference of integer multiples of input frequencies. This property is a principle of frequency converter operation. Typically mixers are three-port devices, with two inputs and one output as shown in Fig. 4.1.



4.1.1 Ideal mixer

From mathematical point of view, the most basic operation that combines frequencies of input signals is multiplication. When the input signals are:

$$V_{RF} = A_{RF} \cos(\omega_{RF} t + \varphi_{RF}) \quad \text{Eq. 4.1}$$

$$V_{LO} = A_{LO} \cos(\omega_{LO} t + \varphi_{LO}) \quad \text{Eq. 4.2}$$

The output is given by:

$$V_{IF} = V_{RF} V_{LO} = A_{RF} \cos(\omega_{RF} t + \varphi_{RF}) A_{LO} \cos(\omega_{LO} t + \varphi_{LO}) \quad \text{Eq. 4.3}$$

Using trigonometric identities:

$$V_{IF} = \frac{A_{RF}A_{LO}}{2} \{ \cos[(\omega_{RF} - \omega_{LO})t + (\varphi_{RF} - \varphi_{LO})] + \cos[(\omega_{RF} + \omega_{LO})t + (\varphi_{RF} + \varphi_{LO})] \} \quad \text{Eq. 4.4}$$

The spectrum of output signal consists of two peaks at frequencies which are sum and difference of the input frequencies (Fig. 4.2). Depending on application, whether it is down or up conversion, one of the signals should be removed by filters. When the LO signal does not change, the amplitude and phase of output signal are proportional to those of the input signal. This feature is very useful as it allows transposing the basic properties of high frequency signal to a lower frequency.

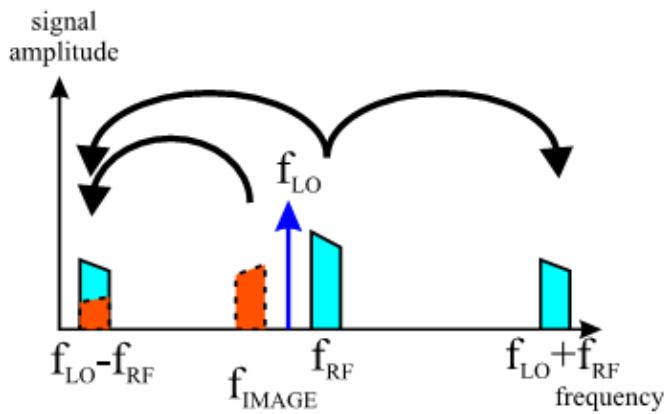


Fig. 4.2 Spectrum of frequency converted signal [14].

4.1.1.1 Image rejection

Signals at frequencies equal to $f_{RF} = f_{LO} + f_{IF}$ and $f_{RF} = f_{LO} - f_{IF}$ both generate a signal at the same intermediate frequency. The given system is designed to work for one of these bands. The other one is called the “image” or unwanted sideband and measures must be taken to prevent it from perturbing the overall operation [15].

In principle there are two possibilities of rejecting this signal: the input (or output, depending on the type of frequency conversion) signal can be band-pass filtered or an image rejection mixer can be employed. Such a device damps one of the signals according to the phase. Note that image and RF signals are converted to the same frequency but they are of different phases. If the input signal is split into two signals in quadrature and then converted with LO signals in quadrature, then one of the sidebands is damped. In the design developed within the framework of this thesis an up conversion as a single sideband

modulator has been used. Its operation can be explained with Fig. 4.3 and following equations:

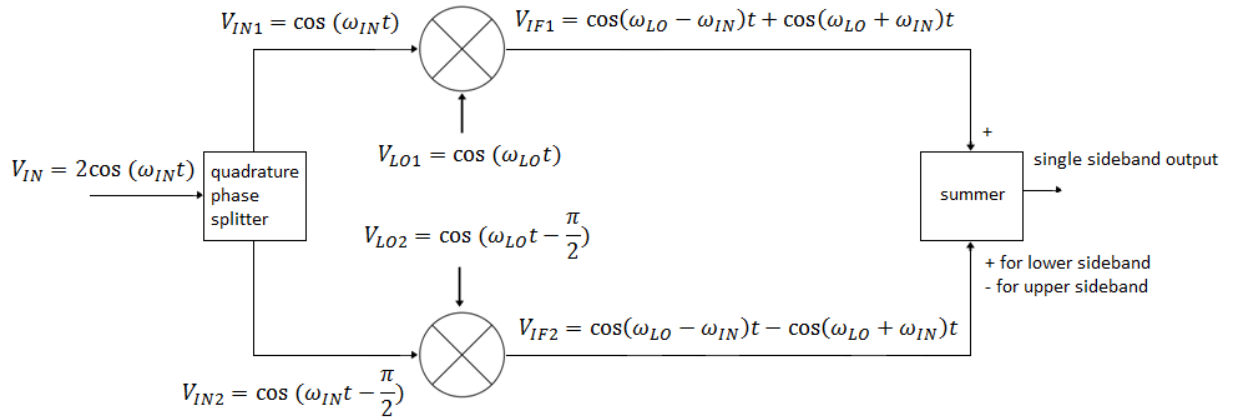


Fig. 4.3 Single sideband modulator.

Input signal is:

$$V_{IN} = 2\cos(\omega_{IN}t) \quad \text{Eq. 4.5}$$

After splitting in quadrature the input signal becomes V_{IN1} and V_{IN2} :

$$V_{IN1} = \cos(\omega_{IN}t) \quad \text{Eq. 4.6}$$

$$V_{IN2} = \cos(\omega_{IN}t - \frac{\pi}{2}) \quad \text{Eq. 4.7}$$

The input signals are mixed with local oscillator signal in quadrature, accordingly:

$$V_{LO1} = \cos(\omega_{LO}t) \quad \text{Eq. 4.8}$$

$$V_{LO2} = \cos(\omega_{LO}t - \frac{\pi}{2}) \quad \text{Eq. 4.9}$$

Signals after mixing are given by:

$$V_{IF1} = \cos(\omega_{LO} - \omega_{IN})t + \cos(\omega_{LO} + \omega_{IN})t \quad \text{Eq. 4.10}$$

$$V_{IF2} = \cos\left(\omega_{LO} - \frac{\pi}{2} - \omega_{IN} + \frac{\pi}{2}\right)t + \cos\left(\omega_{LO} - \frac{\pi}{2} + \omega_{IN} - \frac{\pi}{2}\right)t \quad \text{Eq. 4.11}$$

This result can be directly simplified to:

$$V_{IF1} = \cos(\omega_{LO} - \omega_{IN})t + \cos(\omega_{LO} + \omega_{IN})t \quad \text{Eq. 4.12}$$

$$V_{IF2} = \cos(\omega_{LO} - \omega_{IN})t - \cos(\omega_{LO} + \omega_{IN})t \quad \text{Eq. 4.13}$$

One of the sidebands can be chosen by either adding or subtracting V_{IF1} and V_{IF2} . The same reasoning can be performed to explain the operation of image rejection mixer in down-converting configuration.

4.1.2 Unbalanced mixer

The simplest practical realisation of a mixer is diode. A diode is a non-linear device and its current-voltage characteristic is shown in Fig. 4.5.

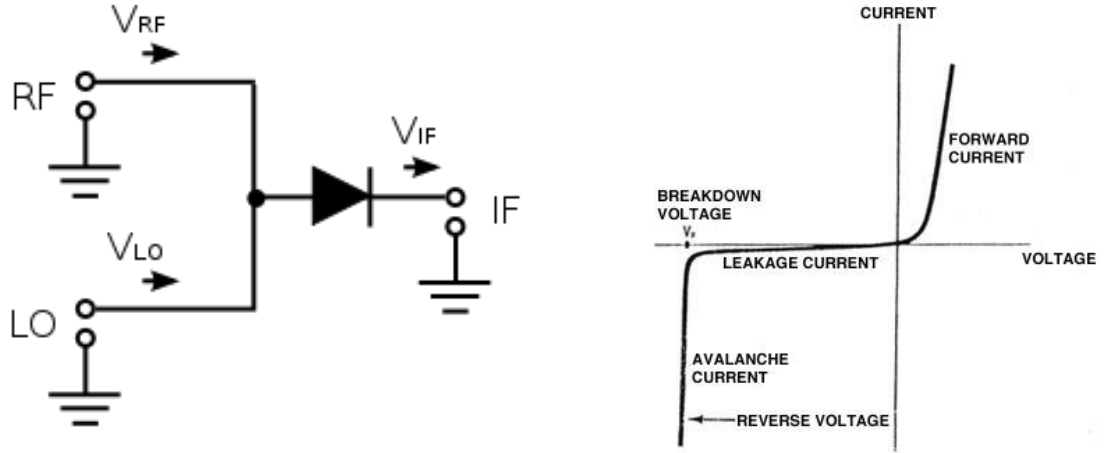


Fig. 4.4 Schematic diagram of diode mixer [16]. Fig. 4.5 Diode current-voltage characteristics [17]. For voltages higher than breakdown voltage a diode current can be described with a power function:

$$i = A \cdot v + B \cdot v^2 \quad \text{Eq. 4.14}$$

When a diode is driven by two sine signals described by Eq. 4.1 and Eq. 4.2 the output is:

$$i = A \cdot [A_{RF} \cos(\omega_{RF}t + \varphi_{RF}) + A_{LO} \cos(\omega_{LO}t + \varphi_{LO})] + B \cdot [A_{RF} \cos(\omega_{RF}t + \varphi_{RF}) + A_{LO} \cos(\omega_{LO}t + \varphi_{LO})]^2 \quad \text{Eq. 4.15}$$

This can be rewritten as:

$$i = A \cdot A_{RF} \cos(\omega_{RF}t + \varphi_{RF}) + A \cdot A_{LO} \cos(\omega_{LO}t + \varphi_{LO}) + B \frac{A_{RF}^2}{2} + B \frac{A_{LO}^2}{2} + B \frac{A_{RF}^2}{2} \cos(2\omega_{RF}t + 2\varphi_{RF}) + B \frac{A_{LO}^2}{2} \cos(2\omega_{LO}t + 2\varphi_{LO}) + BA_{RF}A_{LO} \{ \cos[(\omega_{RF} - \omega_{LO})t + (\varphi_{RF} - \varphi_{LO})] + \cos[(\omega_{RF} + \omega_{LO})t + (\varphi_{RF} + \varphi_{LO})] \} \quad \text{Eq. 4.16}$$

Eq. 4.16 shows that output signal includes input signals, their 2nd order harmonics, DC components and sum and difference of input frequencies. Typically only the last signal is useful, while the rest is distortion.

Very similar to a diode mixer is a configuration with transistor which offers gain of output signal. In case of BJT, the principle of operation is a non-linearity of base-emitter junction. Input signals are applied to the junction and the output signal is a collector current [18].

Unbalanced mixers have strong, not attenuated, input signal, LO signal and higher order intermodulation products at their outputs. Additionally there is also little isolation between each of three ports resulting in undesired signal interactions and feed through to another port [19]. These disadvantages are minimised in balanced mixers.

4.1.3 Single-balanced mixers

Single-balanced mixers consisting of two non-linear devices attenuate LO or RF (but not both) frequency in the output. Its schematic diagram is shown in Fig. 4.6.

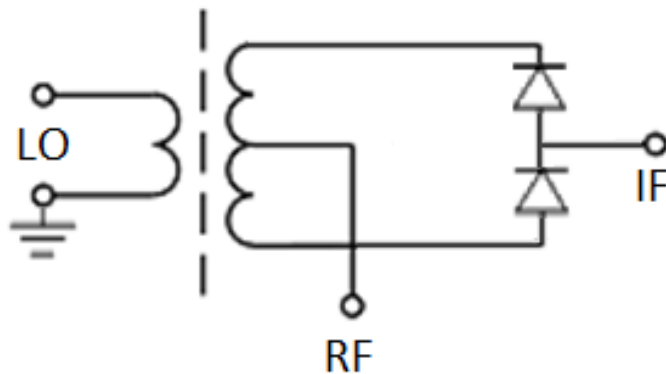


Fig. 4.6 Two-diode single-balanced mixer.

For proper operation the LO signal level needs to be high enough to switch diodes from ON to OFF states. As a result the RF signal is transmitted to IF port only at the instant of high LO level, when the diodes conduct. RF signal is keyed according to LO signal. Spectrum of the output signal consists of sum and difference frequencies and original input frequency. When baluns are perfectly balanced and diodes are matched, the LO frequency is suppressed at the output. Not only the LO port is isolated from IF, but there is also isolation between RF and LO ports. Single-balanced mixers offer better performance than unbalanced mixers, but they require higher LO power. Moreover their cost is higher due to using additional elements.

4.1.4 Double-balanced mixers

Further improvement can be achieved by adding one more diode pair which results in double-balanced mixer which is shown in Fig. 4.7.

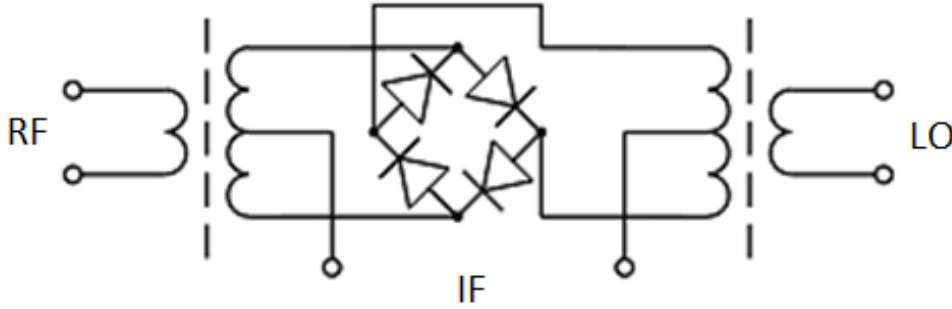


Fig. 4.7 Double-balanced mixer.

In analogy to single-balanced mixers, diode pairs are switched between conducting and open states. The difference is that when the LO signal is low the second pair of diodes is on and RF signal is also transmitted to IF port, but inverted. The RF signal is multiplied by an LO sign, which means it is amplitude modulated by a square wave at LO frequency. This is described by following equations:

$$V_{RF} = A_{RF} \cos(\omega_{RF} t) \quad \text{Eq. 4.17}$$

$$V_{LO} = \text{sgn}[\cos(\omega_{LO} t)] \quad \text{Eq. 4.18}$$

The output is a multiplication of V_{RF} and V_{LO} :

$$V_{IF} = A_{RF} \cos(\omega_{RF} t) \cdot \text{sgn}[\cos(\omega_{LO} t)] \quad \text{Eq. 4.19}$$

The LO signal can be written using Fourier expansion and then IF signal is:

$$V_{IF} = \frac{2}{\pi} A_{RF} \left\{ \cos(\omega_{LO} - \omega_{RF})t + \frac{1}{3} \cos[(3\omega_{LO} - \omega_{RF})t] \dots \right. \\ \left. + \cos(\omega_{LO} + \omega_{RF})t + \frac{1}{3} \cos[(3\omega_{LO} + \omega_{RF})t] \dots \right\} \quad \text{Eq. 4.20}$$

Eq. 2.1 shows that the output spectrum consists only of sum and difference of input frequencies and lower level LO odd harmonics mixing with RF signal. In practical circuits the LO and RF signals can be found at the IF port but strongly attenuated. Moreover, intermodulation products of harmonics are typically easy to filter.

Mixers described in Fig. 4.6 Fig. 4.7 can be built also with transistors instead of diodes.

4.1.5 Gilbert cell

As balanced passive mixers require use of high-frequency baluns they are expensive and cannot be implemented in integrated circuits. Nevertheless there are balanced mixers that are easy in IC implementation. Such a circuit is called a Gilbert cell shown in Fig. 4.8.

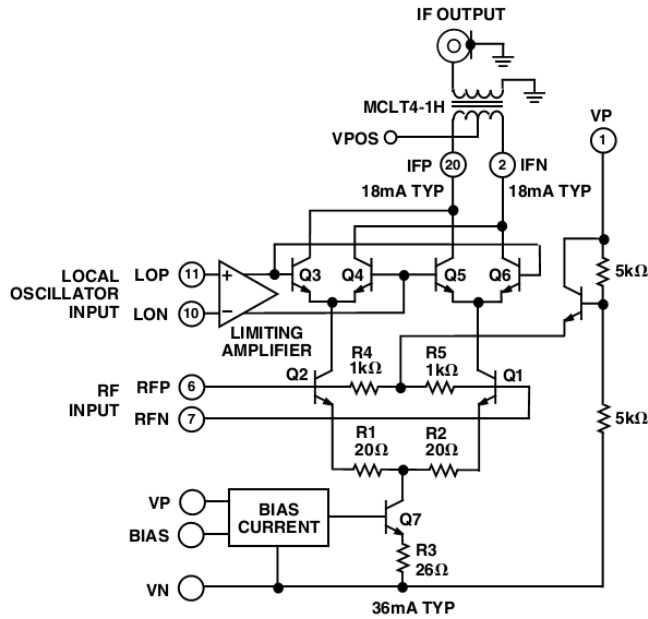


Fig. 4.8 AD831 schematic diagram (configuration with transformer coupling to the IF output) [20].

The Gilbert cell principle of operation is similar to that of passive double-balanced mixer, but baluns are replaced with differential pairs. The RF signal is converted to differential currents by Q1 and Q2 drives the differential pairs Q3, Q4 and Q5, Q6. Limiting amplifier at LO port is used to transform the LO signal to a square wave switching Q3, Q6 and Q4, Q5 between ON and OFF states. It causes keying of RF by LO signal which results in frequency conversion according to Eq. 11.2. The output spectrum consists of sum and difference frequencies of the RF and LO inputs and its lower level products around odd harmonics of LO. RF and LO signals are strongly attenuated in the output.

4.2 Techniques for generating quadrature components

Four basic techniques of generating a quadrature phase shift are described in this section. Three of them are based on filtering methods and phase shift introduced by reactive elements, while the last method employs digital circuit. The description includes principles of operation, advantages and disadvantages in comparison to other methods.

4.2.1 RC-CR phase shifter

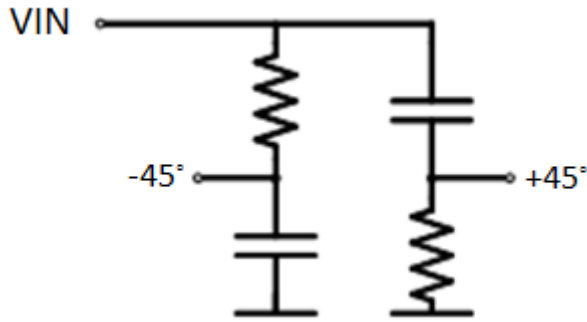


Fig. 4.9 RC-CR quadrature phase shifter.

Fig. 4.9 shows a combination of first-order low-pass and high-pass filters. If the components are of the same values, their phase responses are separated by 90° at all frequencies. However, this simple configuration has disadvantages. Amplitude characteristics of filters are equal only at the pole frequency $f_c = \frac{1}{2\pi RC}$ and are attenuated by only 3 dB. For other frequencies there is an amplitude mismatch between I and Q outputs, limiting the usage only to a narrow frequency range. The amplitude can be normalised by adding limiting amplifiers, transforming the sinusoidal output to a square wave, but it may result in additional mismatch [21; 22]. Furthermore, the circuit is sensitive to tolerances of used components; unequal values of components generate phase error.

4.2.2 Poly-phase filter

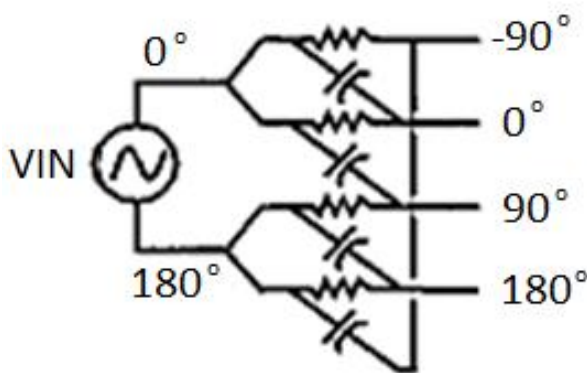


Fig. 4.10 Poly-phase filter.

The poly-phase filter shown in Fig. 4.10 is an evolved version of RC-CR network. Inputs and outputs are symmetrically disposed in relative phases. At the $\frac{1}{2\pi RC}$ frequency the upper input is shifted by $+45^\circ$ and lower by -45° to the output [23]. In practical quadrature shifter realisations the inputs are connected in pairs and driven by differential LO signal.

Consequently outputs are 90° phase shifted. Such a quadrature shifter works only effectively in narrow band. The bandwidth can be increased by adding more poly-phase stages, but it results in larger losses and noise [21]. However, the advantage of the circuit is that it is less sensitive to variations in components values [22].

4.2.3 LC quadrature phase shifter

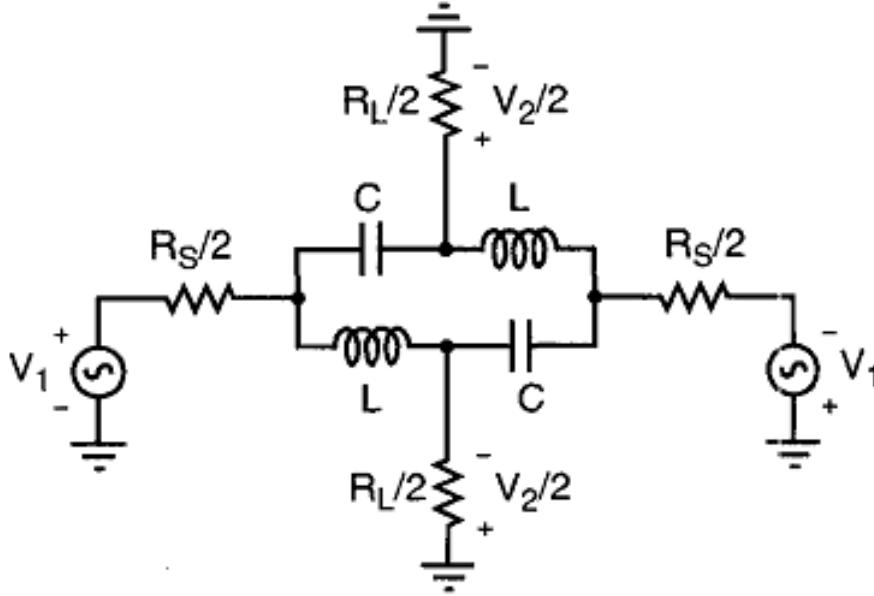


Fig. 4.11 LC quadrature phase shifter.

The circuit shown in Fig. 4.11 is made from a first order LC all-pass filters. At the resonance, when driving signal frequency is equal to $f = \frac{1}{2\pi\sqrt{LC}}$, the output phase is the average of input phases. Driving the circuit in differential mode results in shifting output phase of 90° compared to the input [24]. Opposite to RC-CR network, such phase shifter has almost constant amplitude response, while having very narrow-band quadrature phase characteristics. Since the circuit does not employ resistors it has lower thermal noise and power consumption compared to RC filters.

4.2.4 Divide-by-two method

Previously described circuits use filtering methods to achieve required phase shift. However a quadrature can be achieved in a different way by using a digital circuit consisting of two D-type flip-flops driven in opposite phase. Connecting the $\bar{Q}$ output to the D input results in a change of the output state with every second clock edge. Consequently, the

output frequency (taken from Q output) is divided by two. As the second flip-flop is driven by the inverted signal, its output is 90° phase shifted compared to the first flip-flop. Operation of the circuit is explained in Fig. 4.12.

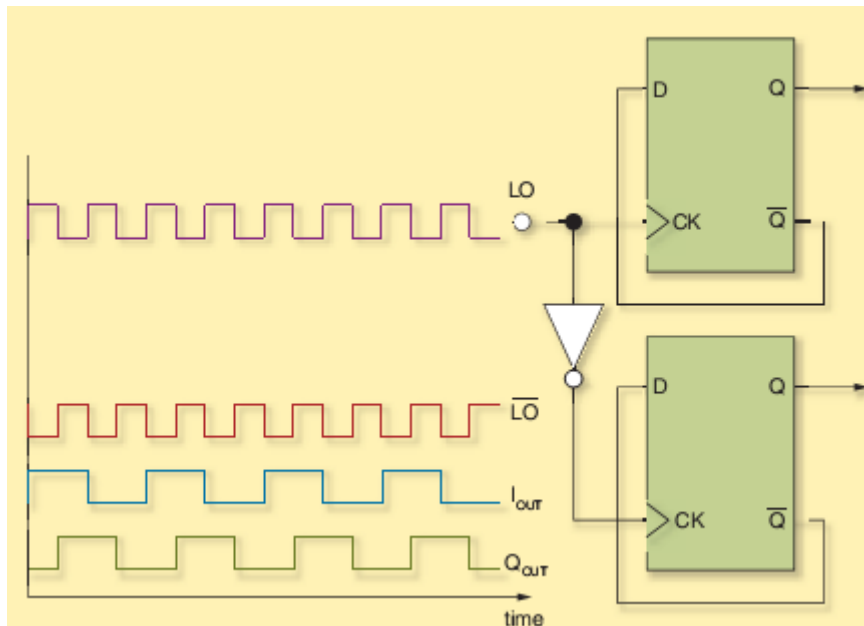


Fig. 4.12 Divide-by-two quadrature generator [25].

Circuit offers a wide-band operation, limited only by the maximum toggling frequency of flip-flops. However it requires driving by a square wave with a very precise 50% duty cycle in order to minimise phase errors [22]. Comparing to previously described methods, the divide-by-two method is more power consuming and requires driving signal at double frequency than the desired one [21].

Chapter 5.

Looking for fundamental components

After understanding the theory, finding required parameters and comparing different techniques the detailed design of the circuit started. First of all it was necessary to choose which components should be used. The most obvious thought was to find an integrated demodulator and modulator. It would result in a simple design consisting of two chips, some passive components and voltage regulators. Moreover, integrated circuits offer outstanding performance as it is easy to create symmetrical, well-matched elements on a silicon chip, resulting in low IQ imbalance.

The frequency converter is needed in feedback modules for two types of cavities: 40 MHz and 80 MHz. As the functionality of the circuit is identical for both modules, it is required to prepare one converter that can work at both frequencies 40 MHz and 80 MHz. Such design will simplify the production process. Consequently, the RF and LO bandwidths should work up to 100 MHz; narrowband circuits cannot be used. An intermediate frequency is chosen to transform the cavities responses to the 10 MHz one, thus for not distorting the signal, the IF 3 dB bandwidth should be well beyond the cavity bandwidths. 20 MHz is enough. As was shown in Section 3.2, the phase imbalance should be less than 2° and amplitude imbalance less than 1.0 dB. All the inputs and outputs should be single-ended, but this is not a strong criterion, as it is easy to convert signals from unbalanced to balanced and vice-versa. Moreover, the noise performance is not crucial, as the input signal is of high level. Knowing the required parameters, the market offers were checked.

5.1 Review of the market

Many manufacturers have quadrature demodulators and modulators, but they are mostly designed for mobile communication systems. Thus, the frequency range does not suit the project requirements. Components satisfying the frequency range criterion are only available from Analog Devices and Linear Technologies. After detailed check it showed up that majority of the circuits have the quadrature generator made with divide-by-two method. As discussed in Section 4.2.4, such method offers a wideband, well matched phase

shift. Furthermore, the need for using the LO driving signal at twice the desired frequency is not a problem, because this signal is available in Proton Synchrotron RF system. However, those circuits did not have the possibility of resetting them. The sign of phase shift generated in divide-by-two method depends on initial conditions, which are the subject to random processes. This situation is unacceptable in the project, because the phase relationship between input and output is crucial in feedback applications. Resetting gives the determined initial conditions, thus solves the problem of unknown phase shift sign. Due to lack of reset circuits, the majority of originally promising chips, like demodulators AD8348 or LT5517 were no longer considered. It was very problematic, because there were no IC demodulators working in such a low frequency range without frequency dividers. Two solutions for this problem were found: either to check if demodulators designed to work in higher frequency range worked also at lower frequencies, or to build the demodulator with discrete components. It was decided to check ADL5380 (400 MHz – 6 GHz) and TRF371109 (300 MHz – 1.7 GHz) demodulators as well as an AD831 active mixer (up to 500 MHz).

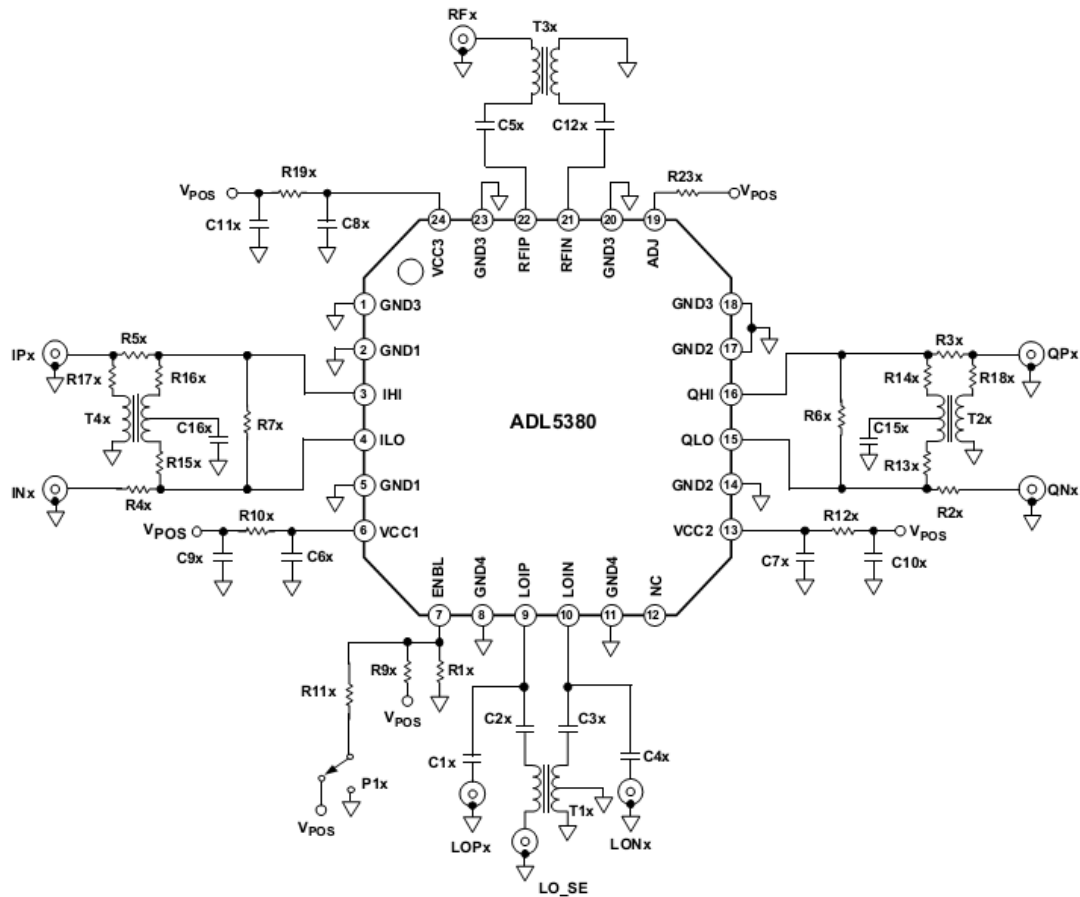
In contrast to problems with the demodulator, there were no doubts with the choice of the modulator. LTC5598 is a quadrature modulator, working with frequencies as low as 5 MHz up to 1600 MHz. The phase shift is achieved with poly-phase filters, thus it can be driven with a sinusoidal signal at the desired frequency. It has also excellent parameters, image rejection about 44 dBc and LO leakage at -55 dBm [26]. Moreover, the output is single-ended and the LO port can be driven either unbalanced or balanced. The LTC5598 has so many advantages, that the decision was obvious – the modulator in quadrature frequency converter will be done with this chip.

5.2 Measurements of evaluation boards

Before using the components in the design, it was decided to order evaluation boards and measure performance of ADL5380 and TRF371109 demodulators, an AD831 active mixer and LTC5598 modulator. Basic parameter of demodulators to measure was the IQ imbalance at RF frequency around 40 MHz and 80 MHz.

5.2.1 ADL5380 demodulator

The ADL5380 is a broadband quadrature IQ demodulator that covers an RF input frequency range from 400 MHz to 6 GHz. The LO interface consists of poly-phase quadrature splitter followed by a limiting amplifier that drives the mixers based on the Gilbert cell design [27]. Its evaluation board (Fig. 5.1) exists in two versions, optimized for operating at lower and higher frequency ranges. Inputs and outputs of the chip are differential, but to simplify measurements, evaluation boards are equipped with baluns. RF and LO input ports provide broadband 50 Ω input impedance and are driven with 1:1 balun to convert differential ports to single-ended. IQ outputs are connected to the 9:1 transformers to provide optimal differential impedance to the chip when the measurements are performed with 50 Ω test equipment. For frequencies lower than 3 GHz, the baluns at RF and LO ports are TC1-1-13. Its frequency range starts from 4.5 MHz, thus there is no need to modify the board prior to measurements at 40 MHz and 80 MHz.



NOTES

1. X = B, FOR LOW FREQUENCY OPERATION UP TO 3GHz, TC1-1-13 BALUN ON RF AND LO PORTS.
X = A, FOR FREQUENCY OPERATION FROM 3GHz TO 4GHz, JOHANSON TECHNOLOGY 3600BL14M050 BALUN ON RF AND LO PORTS.
2. FOR OPERATION BETWEEN 4.9GHz TO 6GHz, THE JOHANSON TECHNOLOGY 5400BL15K050 BALUN, WHICH SHARES A SIMILAR FOOTPRINT AS THE 4GHz BALUN, CAN BE USED.

07556-008

Fig. 5.1 ADL5380 evaluation board schematic.

Measurements of the board at its standard frequency range were started to check if it worked properly. As this board is optimised to work up to 3 GHz, and lowest RF frequency of the chip is 400 MHz, the IQ imbalance was measured at RF frequencies equal to 551.4 MHz, 200.5 MHz, 80.7 MHz and 40.3 MHz. LO frequency was chosen such that the operation at lower sideband and upper was checked with the IF frequency ranging from several hundreds of kHz to about 12 MHz. Frequency of LO generator was changed with a step of 1 MHz. Results are listed in Tab. 5.1.

RF 551.4 MHz			RF 200.5 MHz			RF 80.7 MHz			RF 40.3 MHz		
Imbalance			Imbalance			Imbalance			Imbalance		
LO f [MHz]	Amp [dB]	Phase [°]	LO f [MHz]	Amp [dB]	Phase [°]	LO f [MHz]	Amp [dB]	Phase [°]	LO f [MHz]	Amp [dB]	Phase [°]
548	0.045	0.4	197	0.04	24.95	77	0.04	58.8	37	0.045	72.4
549	0.055	0.15	198	0.07	24.9	78	0.045	58.35	38	0.04	72.05
550	0.055	0.2	199	0.045	25.6	79	0.23	60.2	39	0.04	71.65
551	0.15	1.95	200	0.03	24.95	80	0.355	57.15	40	0.03	71.25
552	0.675	4.5	201	0.035	24.8	81	0.59	60.35	41	0.03	70.85
553	0.195	2.65	202	0.03	24.35	82	0.04	57.55	42	0.04	71.9
554	0.055	0.4	203	0.065	27	83	0.055	57.4	43	0.02	69.95
555	0.035	0.45	204	0.05	23.85	84	0.045	56.85	44	0.01	69.55
556	0.045	0.35	205	0.04	23.65	85	0.04	56.4	45	0.05	70.8
557	0.045	0.4	206	0.045	23.8	86	0.045	56.05	46	0.02	68.6
558	0.04	0.4	207	0.045	23.4	87	0.04	55.7	47	0.03	68.15
559	0.06	0.45	208	0.025	23.1	88	0.065	56.35	48	0.045	67.7
560	0.045	0.4	209	0.04	22.85	89	0.045	55	49	0.06	71.75
561	0.04	0.4	210	0.035	22.7	90	0.04	54.65	50	0.085	66.9
562	0.04	0.3	211	0.045	22.4	91	0.05	54.3	51	0.14	66.4
563	0.035	0.35	212	0.04	22.25	92	0.075	54.6	52	0.115	66.65

Tab. 5.1 IQ imbalance of ADL5380

The results of measurements performed at RF frequency equal to 551.4 MHz show that the evaluation board worked properly. Phase error is typically lower than 0.5° , amplitude imbalance lower than 0.1 dB. Thus, the performance of the ADL5380 at its optimal frequency range is very good, as it is stated in the datasheet. Lowering the RF frequency, well below the preferred range, resulted in increased phase error, while the amplitude balance remained at acceptable level. At an RF frequency equal to 80.7 MHz, the phase error was about 60° and at 40.3 MHz it was about 70° .

The phase error was enormous and even though the amplitude imbalance was at satisfactory level, it became clear that ADL5380 could not be used in the design.

5.2.2 TRF371109 demodulator

TRF371109 is a down conversion receiver from Texas Instruments. It integrates balanced I and Q mixers, LO buffers and phase splitter to convert an RF signal directly to I and Q signals [28]. Moreover it includes programmable output amplifier and low-pass filters. Its frequency range is from 300 MHz to 1700 MHz.

The evaluation board is complex. Inputs and outputs are differential, but the board comes with proper transformers to perform single-ended measurements with 50 Ω instruments. It is possible to bypass them and measure the board in differential mode. The board comes with USB port, via which it can be connected to the computer. Many parameters e.g. output amplifier gain or filters bandwidth can be programmed with a software that evaluation board package comes with.

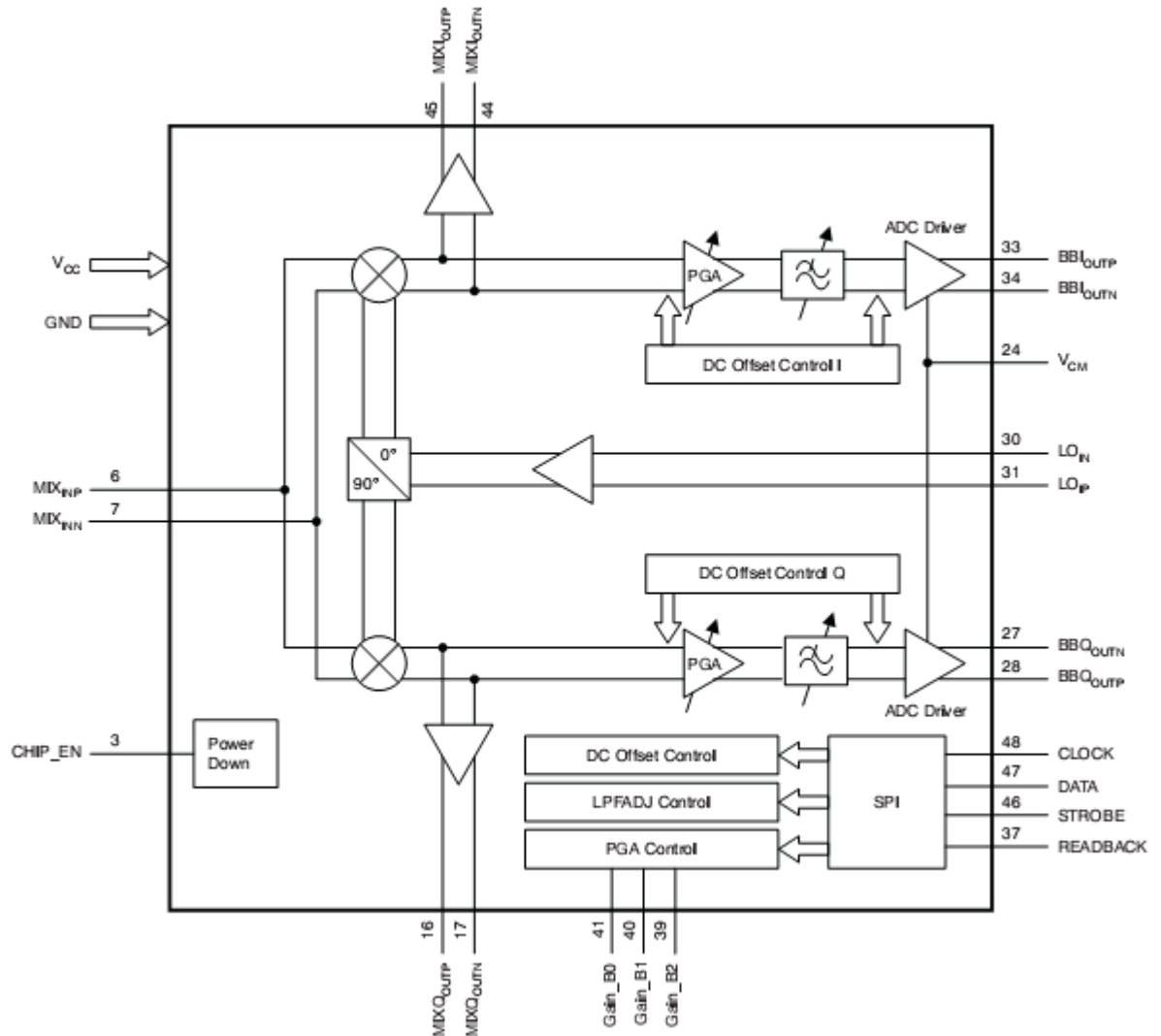


Fig. 5.2 TRF371109 functional diagram [28].

Before measuring the evaluation board the transformers had to be replaced with a different type operating at lower frequencies. Moreover, the decoupling capacitors had to be replaced with the ones of larger capacitances. After applying the modifications the IQ imbalance was measured. The procedure was analogical to measuring the ADL5380: starting with a standard frequency range and then measuring at the project range. Tab. 5.2 shows the results of TRF371109 measurements.

RF 551 MHz			RF 39.725 MHz		
	Imbalance			Imbalance	
LO f [MHz]	Amp [dB]	Phase [°]	LO f [MHz]	Amp [dB]	Phase [°]
548	0.2	0.75	37	0.855	10
549	0.2	0.5	38	1.005	10.5
552	0.205	1.75	42	1.385	14
553	0.195	0.15	43	1.42	14.5
554	0.2	0.3	44	1.435	15.1
555	0.195	0.2	45	1.45	15.55
556	0.205	0.4	46	1.44	15.95
557	0.19	0.4	47	1.425	16.35
558	0.19	0.5	48	1.385	16.65
559	0.185	0.65	49	1.33	16.8
560	0.19	0.7	50	1.26	16.85
561	0.185	0.85	51	1.185	16.7
562	0.175	0.9	52	1.12	16.35

Tab. 5.2 IQ imbalance of TRF371109

Even the modified evaluation board of TRF371109 has an acceptable performance at 551 MHz. However, this is not the case at the design frequency range. At around 40 MHz the imbalance is quite big: about 1.0 dB in amplitude and about 15° in phase. Amplitude mismatch could be compensated using an attenuator, but it is hard to compensate the phase error. The I output signal contained additional peaks in the spectrum, compared to Q output, as shown in the Fig. 5.3. In addition, while increasing the input signal level, the I output saturated before the Q output.

Not only due to the IQ imbalance, but also because of the unreliable behaviour of the chip in the input low frequency range, the TRF371109 could not be used in the design.

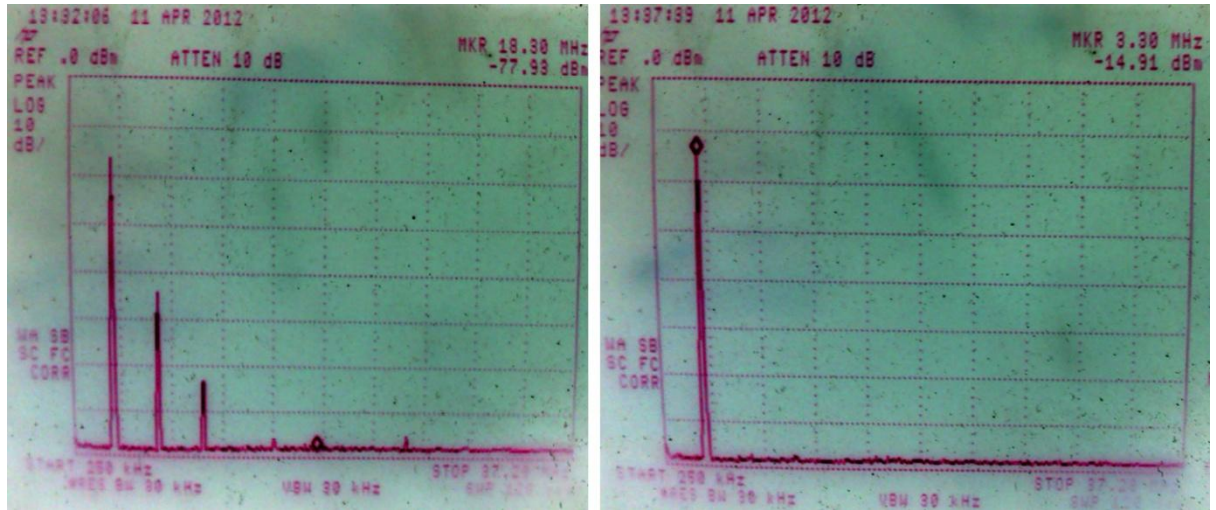


Fig. 5.3 TRF371109 I (on the left) and Q (on the right) output spectrums at RF = 39.725 MHz.

5.2.3 AD831 active mixer

As the ADL5380 and TRF371109 integrated demodulators could not be used in the project, the demodulator had to be built with single mixers. The AD831 mixer was chosen, because of the following favourable properties:

- Frequency range that allows the use in the project;
- integrated output amplifier that could be used for conversion loss compensation;
- low distortion;
- high isolation between the ports;
- single-ended inputs and outputs.

The goal of measurements was to find the optimal input signal parameters, especially considering the distortion level and voltage gain. According to the AD831 datasheet, the optimal LO drive level is -10 dBm [20]. However, increasing the LO level to 0 dBm results in lowering the unwanted spectral components. It is shown in Fig. 5.4. The Y axis is a level in dBc with the difference frequency chosen as a reference. All the peaks despite those at the sum and difference frequencies of RF and LO signals are distortion. As the green line (a spectrum for a 0 dBm LO level) is below the other lines, it is better to drive the LO port at 0 dBm, especially that LO leakage is constant and equal to around -30 dBm. The chart was made for the worst case from the distortion point of view, which is upper sideband frequency conversion with detuned cavity. The detuned cavity lowers the frequency of the strongest peak, which in addition to upper sideband frequency results in distortion closer to the signal wanted. Thus, the filtering task is harder. The closest important distortion peak is

at 19.2 MHz which is the $2LO - RF$ frequency. If the lower sideband conversion is chosen (LO frequency higher than RF), the distortion is at higher frequency. Therefore a filtering is much easier. Additional advantage of driving the LO port with signal at 0 dBm is focusing the voltage gain around a constant value for different RF levels and frequencies as shown in Fig. 5.5. Set 1 means RF frequency 36.6 MHz, LO 27.9 MHz and set 12 is RF 80 MHz, LO 89.5 MHz.

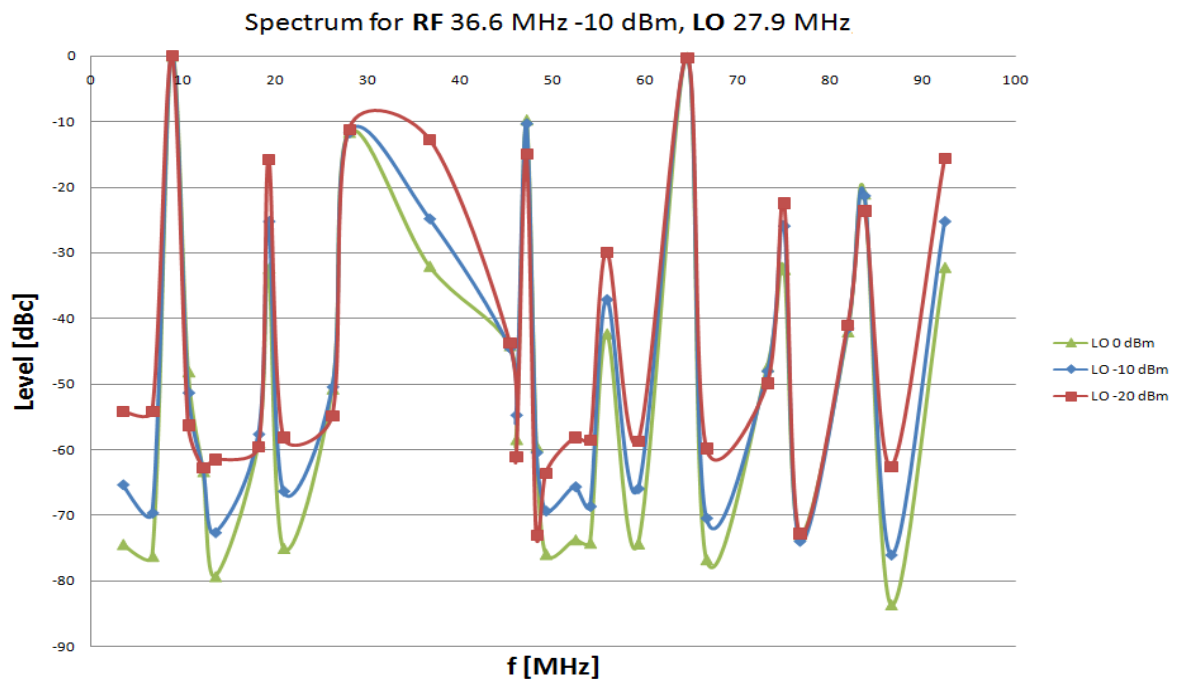


Fig. 5.4 Influence of the LO drive level on the distortion in AD831.

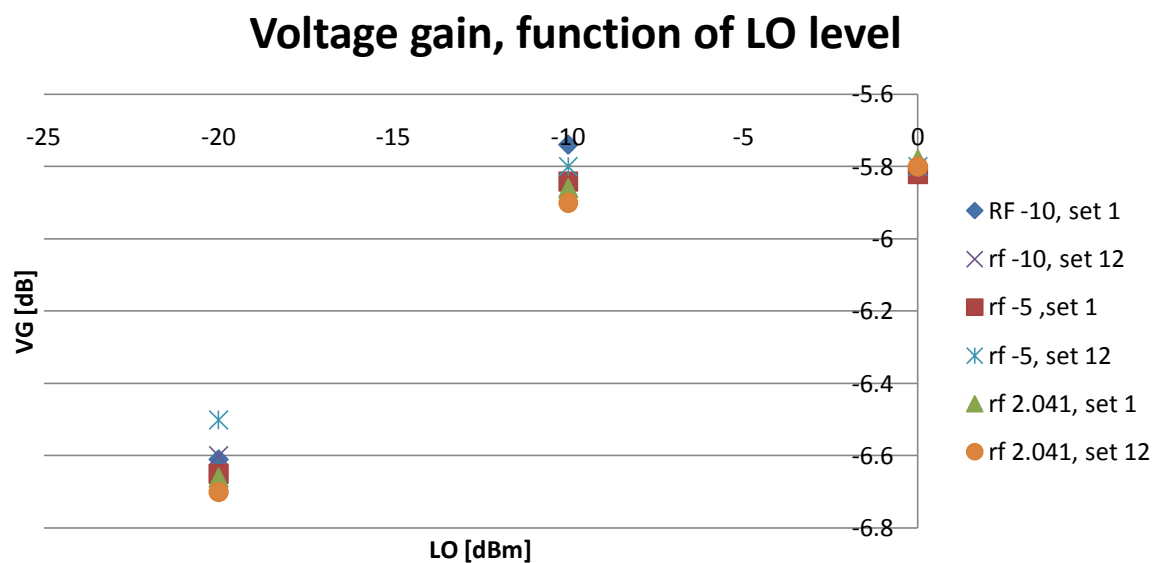


Fig. 5.5 AD831 voltage gain vs. LO level.

The performance of the chip was satisfactory, thus it was chosen to build the demodulator.

5.2.4 LTC5598 modulator

LTC5598 modulator is a direct quadrature modulator designed for wireless applications such as GSM, EDGE or CDMA. It may be also configured as an image rejection up-converting mixer, by applying 90° phase shifted signals to the I and Q inputs [26]. In such configuration it was used in the design. The LTC5598 frequency range is from 5 MHz to 1600 MHz, at frequencies between 40 MHz and 80 MHz a voltage gain is around -2.3 dB, image rejection -45 dBc and LO leakage -55 dBm. With the evaluation board it was decided to check the performance.

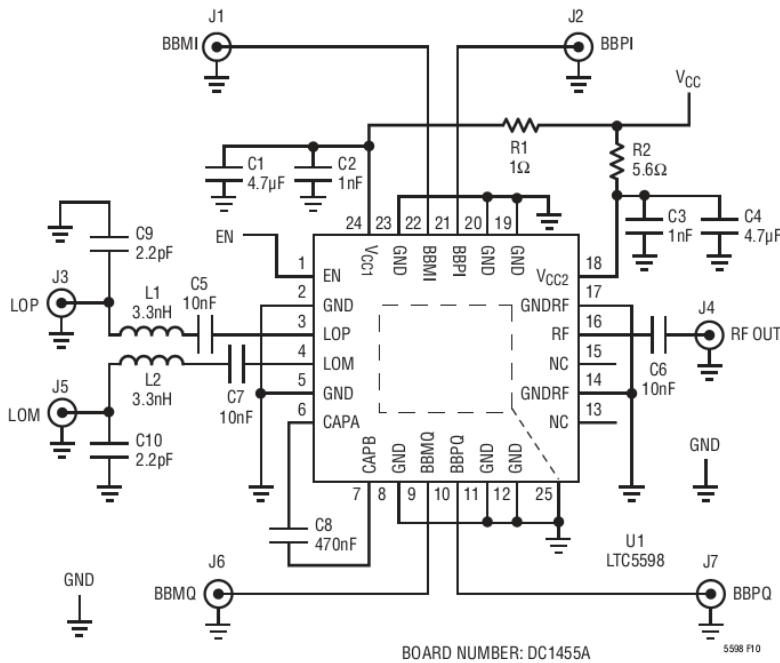


Fig. 5.6 LTC5598 evaluation board schematic [26].

Prior to measurements the board was modified. Firstly, the board can be optimised to three frequency ranges by means of LO input matching. The frequency ranges are:

- 5 MHz to 250 MHz;
- 80 MHz to 1300 MHz;
- 1150 MHz to 1600 MHz.

Matching is done by changing the values of passive components at LOP and LOM inputs (Fig. 5.6) values according to Table 2 in [29]. Originally the board was optimised for the middle frequency range. To optimise the board for lower frequencies, C5, C7 were replaced with 100 nF capacitors, C9, C10 with 120 Ω resistors and L1, L2 with 0 Ω resistor. Moreover, as the baseband inputs are differential and require DC bias, transformers had to be added with centre tap connected to a DC power supply.

Measurements showed, that all the parameters are practically the same as in the datasheet. Although the 0 dBm LO drive level is suggested, the performance for two more values, -10 dBm and 10 dBm was investigated. Voltage gain (Fig. 5.7) and image rejection (Fig. 5.8) were measured for a couple of baseband amplitudes and frequencies.

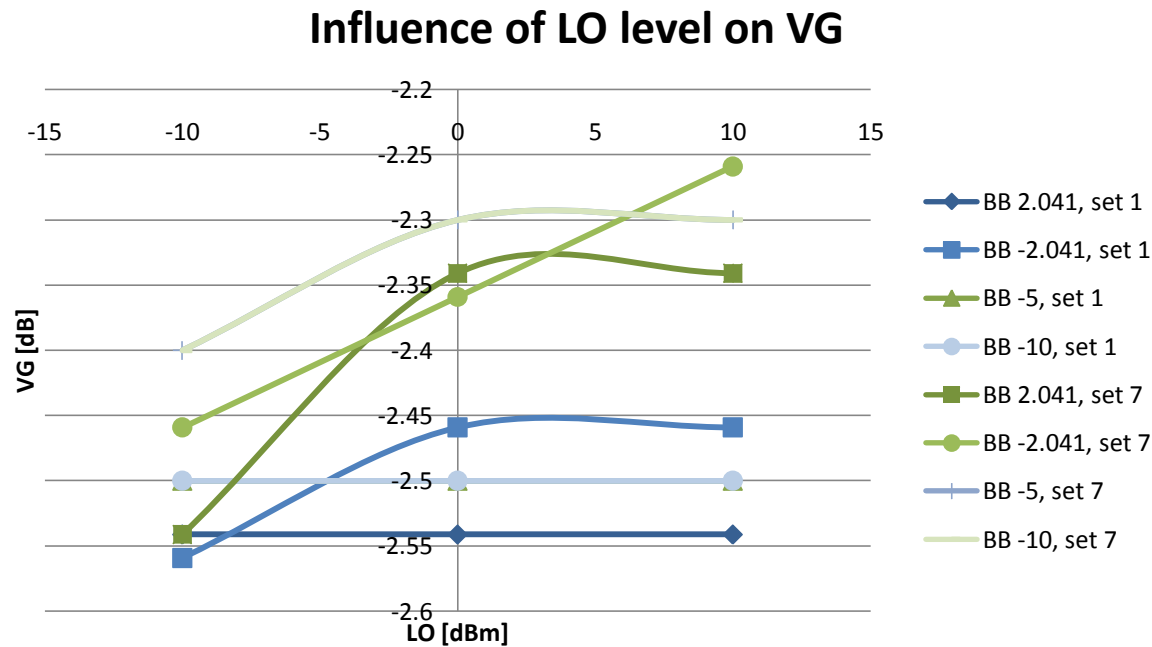


Fig. 5.7 LTC5598 voltage gain vs. LO drive level.

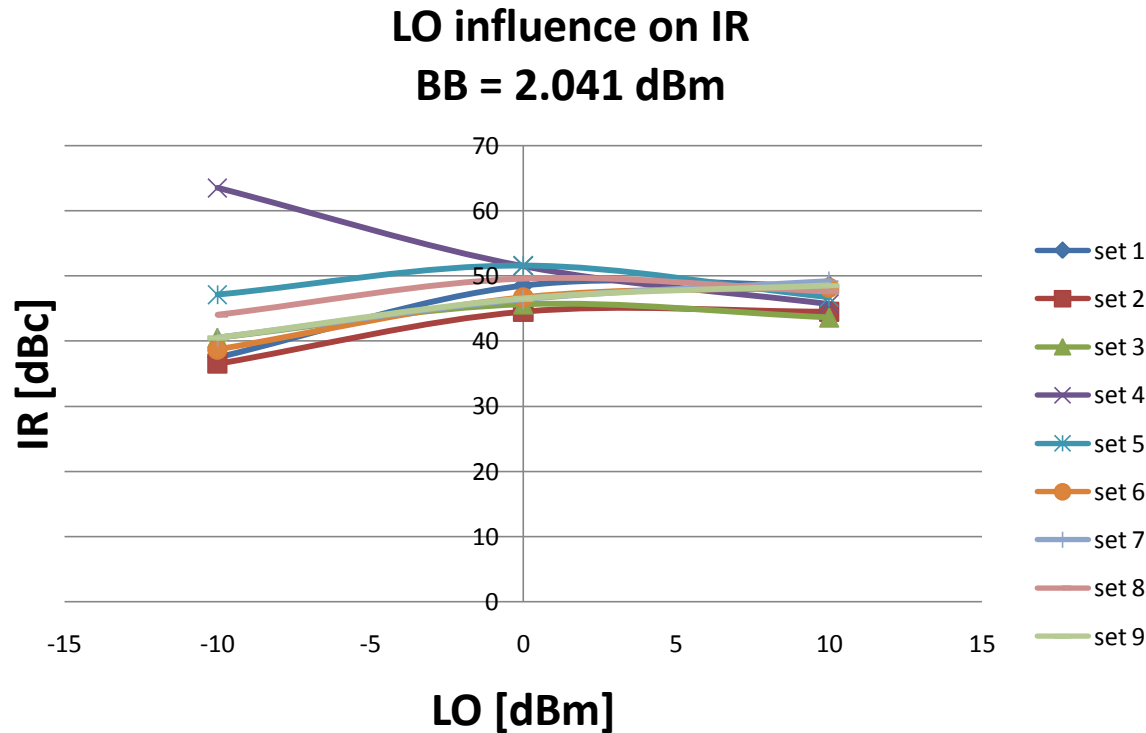


Fig. 5.8 LTC5598 image rejection vs. LO drive level.

Lowering the drive level is usually required as it results in lowering the electromagnetic interference. However, the parameters of both the LTC5598 and AD831 were better when LO input was driven with 0 dBm signal.

5.3 Chosen components

As the measurements showed, the ADL5380 and TRF371109 demodulators could not be used in the design, due to their unacceptable IQ imbalance at the project frequency range, that is 40 MHz and 80 MHz. The demodulator had to be done with discrete mixers. AD831, an active mixer from Analog Devices, was chosen. It characterises in low distortion, single-ended inputs and output, built-in amplifier which can be used to compensate for conversion loss. The LTC5598 quadrature modulator has satisfying parameters, thus it was used in the design for the up conversion part of the frequency converter.

The quadrature frequency converter is based on AD831 and LTC5598 chips, but obviously the project does not consist only of them. The detailed look into the design and the additional components is given in the next chapter.

Chapter 6.

Prototype board

Instead of manufacturing the final version of the frequency converter, the prototype board was made first. It can also work in the feedback module, but it has additional elements, such as jumpers that allow to bypass some functions and help in debugging process. The schematics of the prototype board has been designed under my responsibility, but the PCB layout was entrusted to the experienced CERN PCB designers. IQ mismatch, especially the phase error, is sensitive to a symmetry of the circuit. A lot of attention was paid to ensure, that all the traces lengths are equal and that components are located in the same distance to crucial elements.

In this chapter, the frequency converter prototype board is described in detail, including possible adjustments, list of modifications and performance characteristics. The block diagram of the prototype can be found in Fig. 6.1, the picture of the board in Fig. 6.2. Numbers of elements refer to the frequency converter prototype board schematic which can be found in Appendix V.

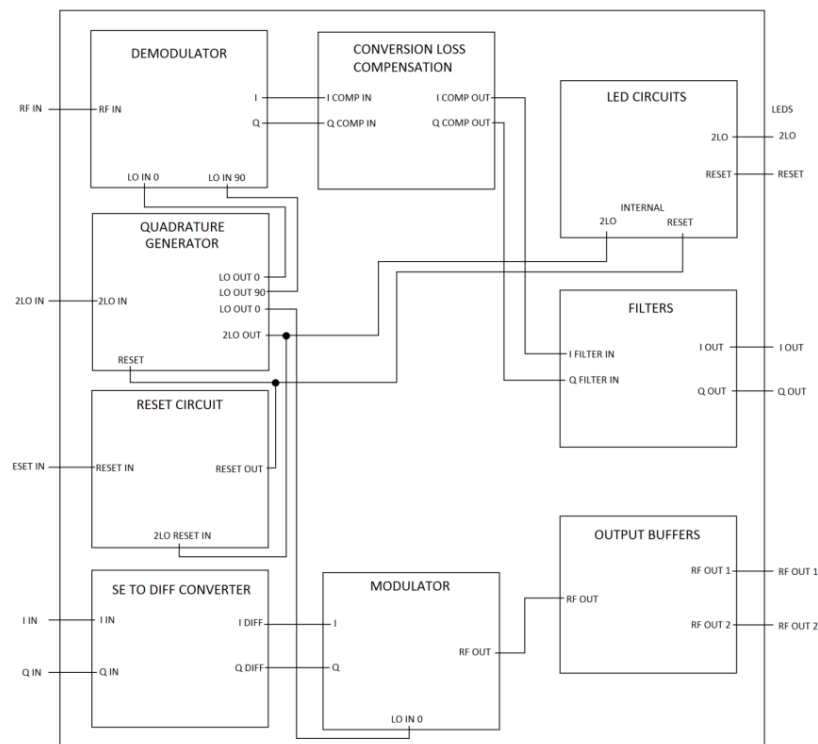


Fig. 6.1 Block diagram of the quadrature frequency converter.

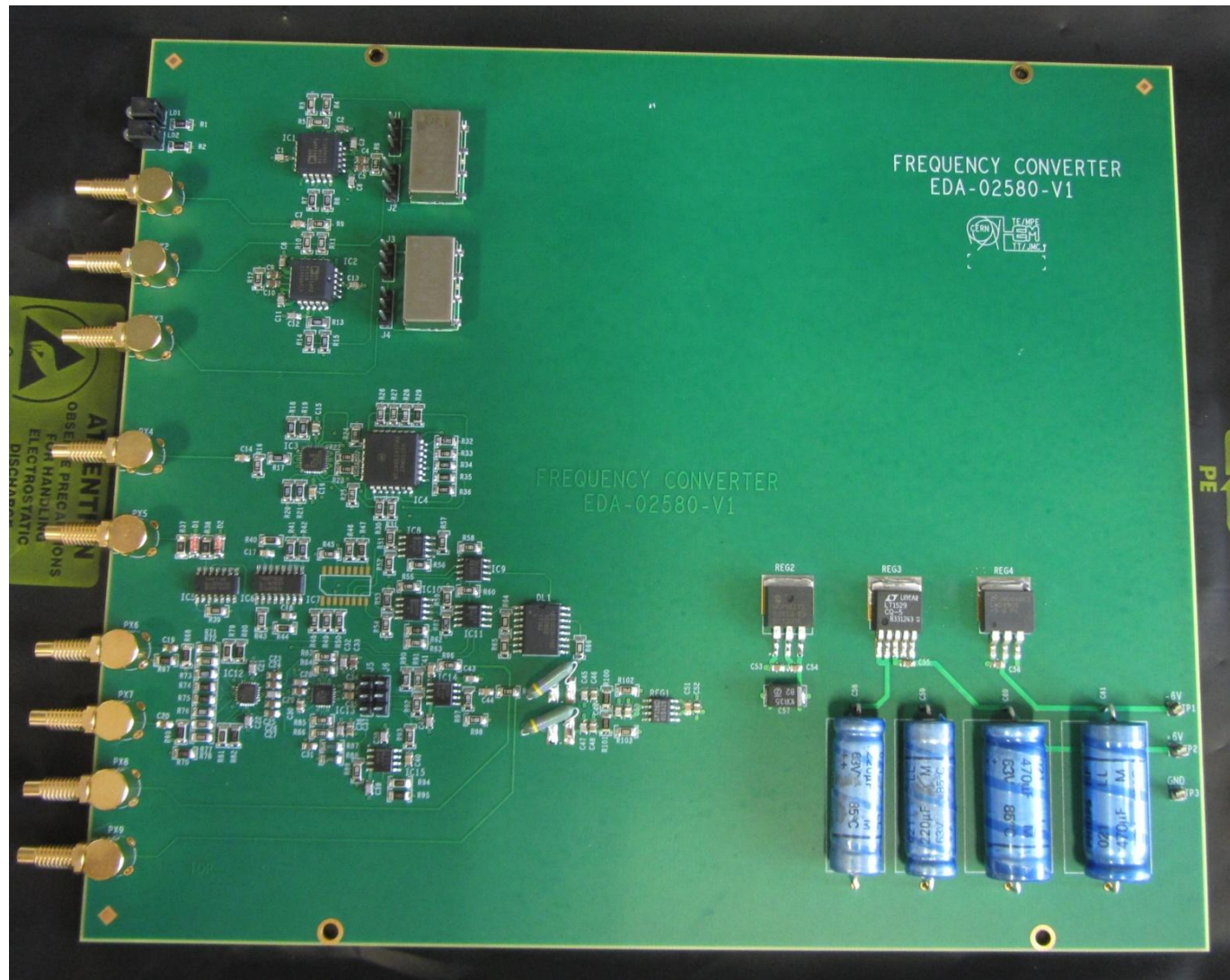


Fig. 6.2 Quadrature frequency converter prototype board.

6.1 Description of the prototype board

The frequency converter is designed to be a part of one-turn delay feedback for CERN Proton Synchrotron high frequency cavities. RF to IF down conversion is done by an IQ demodulator to get in-phase and quadrature components. The IQ modulator is used for up conversion, normally configured as an image rejection up-converting mixer by driving the inputs in quadrature. All the internal signals are sent via 50 Ω transmission lines.

The demodulator is made of two AD831 double balanced active mixers (IC1, IC2). The quadrature generator is based on frequency division by ECL D-type flip-flops (IC4), thus the board has to be fed with LO signal at twice the frequency desired for mixing. The IQ outputs are low-pass filtered (FL1, FL2) with a 3-db cut-off frequency of 24 MHz to eliminate odd harmonics of the LO signal – typical spectrum component of doubly balanced mixers. For best performance, the frequency of IQ signals should be around 10 MHz.

The LO section consists of a comparator (IC3) which transforms the input waveform (typically a sine wave) to the ECL compatible signal. The comparator has two pairs of differential outputs. One differential pair is connected directly to the inputs of a D-type flip-flop to generate in-phase and quadrature LO signals driving the demodulator. The inverted signal is used to generate the in-phase signal; the non-inverted one drives the flip-flop which creates the quadrature signal, at 90°. The in-phase signal is delayed relative to quadrature. The LO quadrature phase shift is identical to the one generated internally in the integrated IQ modulator (IC13) used for up conversion. The second differential output of comparator is used to drive modulator, reset and LED circuits. The inverted signal drives the flip-flop connected to the modulator. The flip-flop output signal is filtered before applying to the modulator, thus there is an additional delay which has to be taken into account when estimating a total delay. The non-inverted signal is buffered to three copies. One is used to drive the LED circuit indicating the existence of LO signal, the other two are used to drive the reset circuit. The LO LED circuit (IC6, IC9, IC10, IC11, DL1, LD1) is quite complicated, as the fast changing ECL signal has to be transformed to drive a TTL monoflop which slows the signal to make it visible.

The flip-flop reset circuit is designed to ensure the proper phase shift of the LO signal in the demodulator. The reset is internally synchronised (by flip-flops IC4 and IC8), thus the external reset pulse can be sent at any time. The input stage of reset circuit is made of

a voltage divider followed by a TTL inverter (IC5) externally protected by clamping diodes. For proper operation the reset pulse should be 10 V in amplitude. The circuit contains also a LED (LD2) to indicate the arrival of the reset pulse which resets the quadrature generator.

The up conversion path consists of an LTC5598 high linearity direct quadrature modulator (IC13). The IQ inputs are connected to differential amplifiers configured as single-ended-to-differential converters (IC12). The differential outputs drive the baseband inputs of the modulator. The baseband inputs have to be DC biased at 0.5 V which is assured by a precision reference LT1019 (REG1). The LO signal is derived from flip-flop, filtered and fed into the modulator. Filtering is required, because internal quadrature generators do not work properly with a square wave. The LTC5598 use doubly balanced mixers, thus the RF output spectrum contains also odd harmonics of the LO signal. The output is not filtered, but directly fed into the cascade of THS3201 (IC15) and THS3202 (IC14) current-feedback amplifiers. The THS3201 is configured to compensate the loss of the IQ modulator; the THS3202 serves as output buffers to drive two outputs.

The power supply section consists of two inputs: 6 V and -6 V requiring about 1.2 A and 0.4 A current respectively. They drive voltage regulators to create stable +/- 5V power supply (REG3, REG4), 3 V ECL termination voltages (REG2) and a precision reference used to DC bias the baseband inputs of modulator (REG1).

All inputs (despite a reset) and outputs are 50 Ω , single-ended. Connectors are SMC.

The board is designed to fit in the VME crate. The aluminium front panel has printed labels describing every connector. The labels are put in sections to easily identify the demodulator and the modulator part of the board. The front panel layout can be found in Appendix IV.

6.2 Adjustments

The jumpers J1, J2, J3, and J4 allow the user to either bypass or connect the low-pass filters on outputs of the demodulator. Resistors R3, R5, R13, and R15 set the gain of mixers output amplifiers. They can be adjusted to compensate for a conversion gain. Note that the mixer output is doubly terminated, thus the set gain has to be twice the desired. To keep the IQ amplitude imbalance low, high-precision resistors should be used at this point.

The jumpers J5 and J6 can bypass IC15, the output amplifier of the modulator which compensates losses during up conversion. The gain setting resistors are R94 and R95. This

amplifier is also doubly terminated. IC15 is a current-feedback amplifier, thus the THS3201 datasheet [30] can be consulted to find the proper resistor values.

IC14 is a dual amplifier configured as doubly terminated buffer. Resistors R90, R91, R97, R98 set the gain. It is also a current-feedback amplifier, for proper resistor values the THS3202 datasheet [31] can be referred to.

6.3 List of modifications following the commissioning of the prototype

As it is usual with prototypes, some modifications had to be done in order to make the frequency converter work properly. They are listed below.

- Added a 100 Ω resistor to inputs of the THS3202 to suppress oscillations.
- Swapped pins 3 and 4 of IC7.
 - The pin 3 is directly terminated to 50 Ω , the pin 4 is connected to the pin 23 of IC4. The goal of this modification was to invert a clock signal for the reset circuit of IC4 flip-flop to ensure a proper phase shift of quadrature generator.
- The MCP1827 3V voltage regulator replaced with the MC7902 -2V negative voltage regulator.
 - The regulator had to be replaced with a negative voltage regulator in order to be capable of sinking a current, because it is used as an ECL termination voltage. The V_{in} pin is connected to a ground plane, the GND pin to a 5V plane and a V_{out} pin to a 3V plane. Because of the fact that the MC7902 is no longer produced, any negative, adjustable voltage regulator can be used. The circuit should be designed in such a way that the output gives -2V voltage relative to the GND pin, the GND pin is connected to a 5V plane and the V_{in} pin is connected to the ground plane. In such configuration the output voltage of the circuit will be 3V relative to ground, capable of sinking the current.
- Added a low pass filter and a 100 nF capacitor between R110 and C29.
 - In order to generate a proper LO phase shift the modulator needs to be driven with a sine wave, thus the ECL signal has to be low pass filtered. Due to the fact that an ECL bias voltage differs from the bias voltage of the

LTC5598 LO input, the filter input and output needs to be DC blocked. 100 nF ceramic capacitors are recommended.

6.4 Measurements

After applying modifications, measurements of the performance of the prototype quadrature frequency converter were made. The most important results are small IQ imbalance of demodulator and the fact, that the whole board acts as a part of a cable – small attenuation with a constant delay (when input signal is applied to RF input, demodulator IQ outputs are directly connected to modulator IQ inputs, and output signal is taken from RF output). Due to the fact, that modulator LO signal has to be filtered and that one board was produced, the board was prepared to work only with 40 MHz cavity. Therefore, the measurements of the board were performed only at this frequency. However, it is only one element that makes a difference between 40 MHz and 80 MHz cavity frequency converter, thus the production process is still simplified – one PCB layout is sufficient. All the characteristics are in Appendix III.

The demodulator down converts signals with a small IQ imbalance. For lower side-band mode the phase error is typically better than $\pm 0.5^\circ$ (Fig. 6.3) amplitude imbalance is around 0.06 dB (Fig. 6.4).

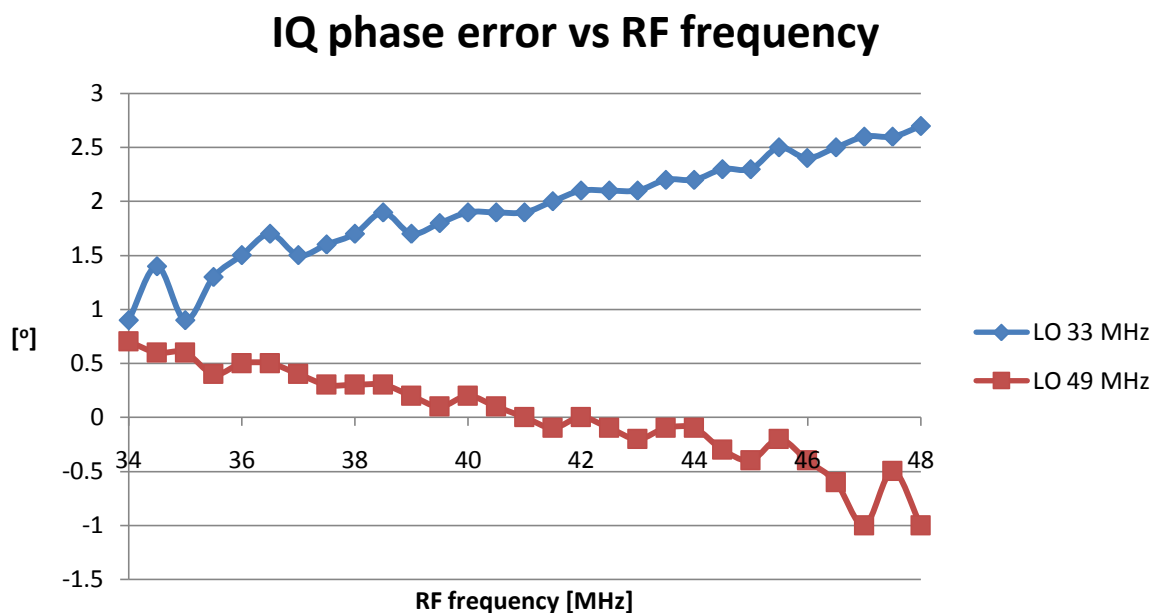


Fig. 6.3 Demodulator phase error. Chart is symmetrical for both sidebands when intermediate frequency is taken into account. Phase error increases with intermediate frequency, thus IQ output filters can be considered as a source of mismatch.

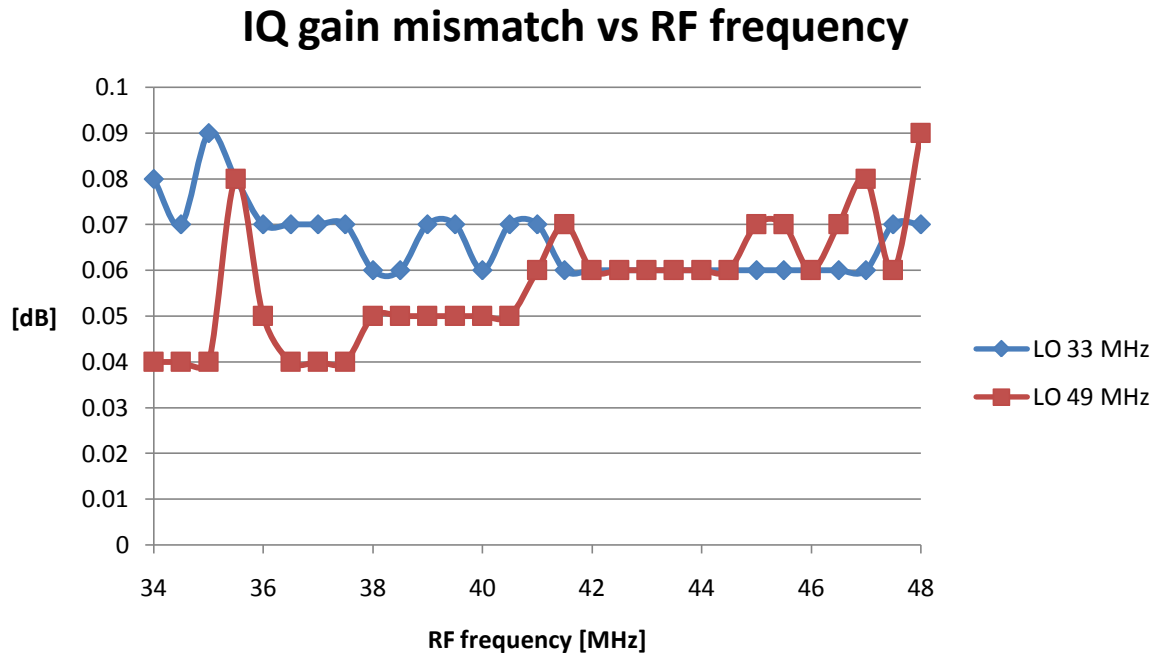


Fig. 6.4 Demodulator amplitude mismatch.

With the demodulator outputs connected directly to the modulator inputs, a transfer function of the frequency converter (Fig. 6.5, Fig. III.3 and Fig. III.4) is flat in gain (mind the gain scale on the chart which is 0.1/0.2 dB per division) and in phase.

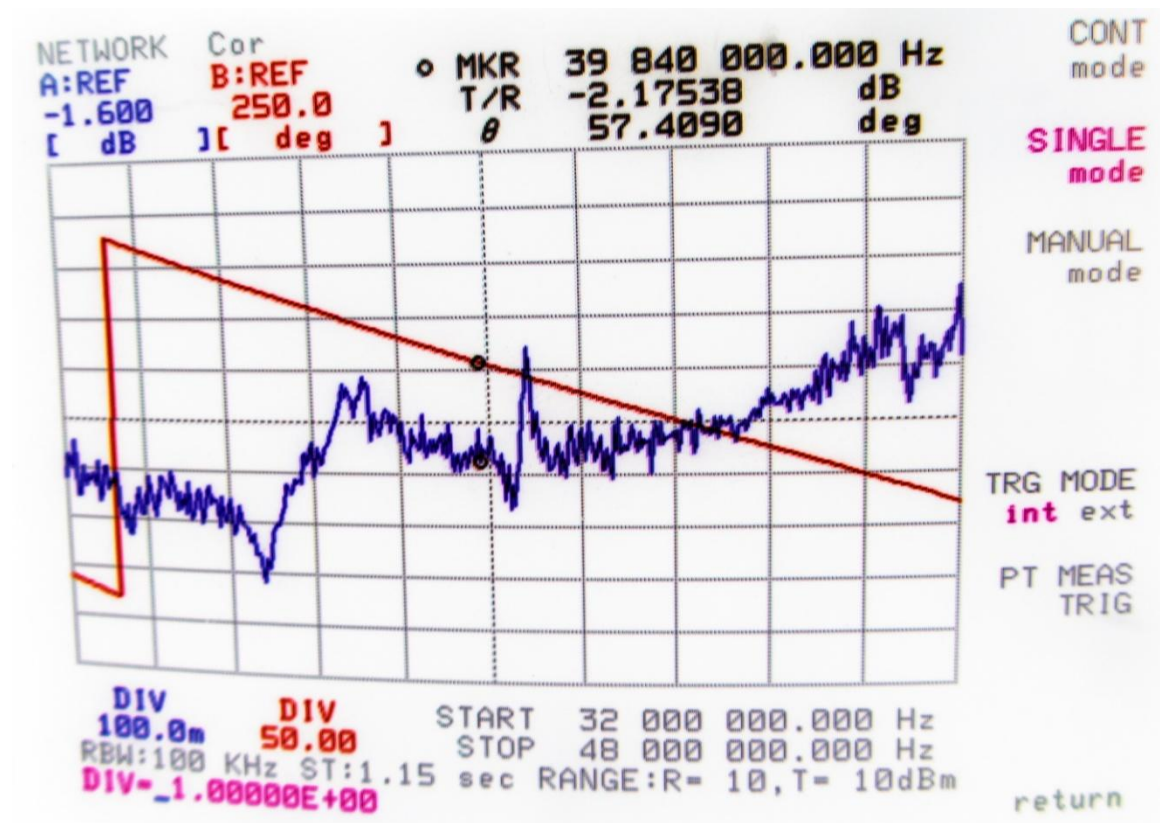


Fig. 6.5 Transfer function, LO frequency 49 MHz (blue trace – gain, red – phase).

Image rejection is typically higher than 28 dB (Fig. III.5). Demodulator conversion gain is about -0.3 dB (Fig. III.6), third order intermodulation products levels are lower than -41 dBc in both up and down conversion parts (Fig. III.7 to Fig. III.10). 1 dB compression point is higher than 10.5 dB over the required RF signal frequency range (Fig. III.11 and Fig. III.12). LO leakage to IQ outputs (Fig. III.13) lower than -35 dBm in a USB mode and lower than -70 dBm in a LSB mode, to RF output is better than -35 dBm (Fig. III.14). An I output to RF input isolation is higher than 50 dB (Fig. III.15), a Q to RF isolation is better than 60 dB (Fig. III.16). A 2LO input return loss is better than 25 dB (Fig. III.17), RF input is better than 20 dB (Fig. III.18). IQ inputs return losses are higher than 30 dB (Fig. III.19 and Fig. III.20).

Chapter 7.

Summary

The objective of the thesis was to develop and build a prototype of the quadrature frequency converter that would work as a part of the one-turn-delay feedback. A signal from a high frequency cavity is down-converted, processed by a digital part of the one-turn-delay feedback module, then up converted back and applied to further elements of the feedback loop. The fact of frequency conversion should be “invisible”, thus the whole part should behave like a cable – small attenuation and constant delay.

To find the required parameters of the converter, the project was started with performing simulations in the Mathematica software. As shown in Chapter 3, the phase error should be less than 2.0° and amplitude mismatch smaller than 1.0 dB. The next step was to check what components are available. To ensure the smallest possible mismatch it was desired to design the converter with integrated demodulator and modulator. As described in Chapter 5 a perfect integrated modulator was found, but the demodulator had to be built with discrete mixers. The divide-by-two method was chosen for the quadrature generator as it offers good mismatch over wide frequency range. In such a configuration, the PCB layout is more flexible, as the board can be used as well in the 40 MHz cavity as in the 80 MHz. After finding all required additional elements a schematic of the frequency converter was created. Afterwards, the PCB layout was designed by a CERN designer. The most important requirement was maximum symmetry in the demodulator. All the traces had to have equal lengths, chips should be at the same distance.

When the board was manufactured, some problems occurred (as usual with prototypes). After applying the modifications to the board its performance was measured. The results were very good, fully satisfying the project requirements. The demodulator IQ amplitude mismatch was around 0.06 dB, while the phase error for lower sideband frequency conversion was better than $\pm 0.5^\circ$. Moreover, the transfer function of the system with the demodulator connected directly to the modulator is flat in gain and phase.

The board will be installed in the Proton Synchrotron RF system once the one-turn-delay feedback module is finished.

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Appendix I

Calculations on IQ mismatch

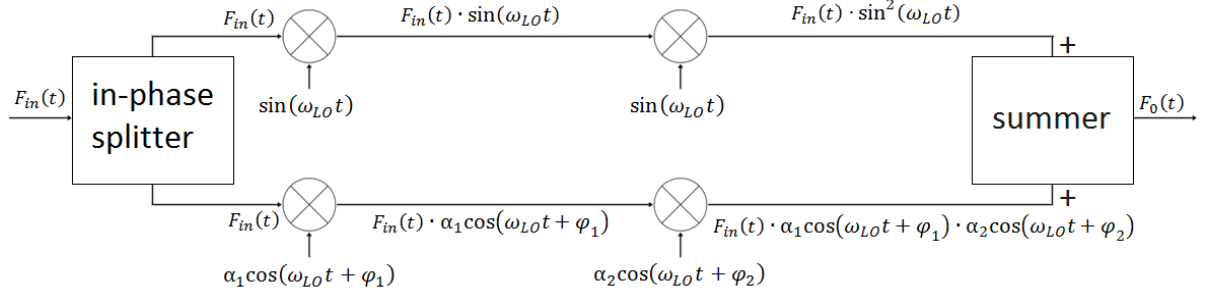


Fig. I.1 Block diagram of double frequency conversion.

An input signal, after passing through the frequency converter (Fig. I.1) may be given by:

$$F_0(t) = [\sin^2(\omega_L t) + \alpha_1 \alpha_2 \cos(\omega_L t + \varphi_1) \cos(\omega_L t + \varphi_2)] F_{in}(t) \quad \text{Eq. I.1}$$

Where:

$F_{in}(t)$ - input signal;

$F_0(t)$ - output signal;

ω_L - angular frequency of a heterodyne signal;

α_1, α_2 - amplitude mismatch introduced by 1st and 2nd frequency conversion;

φ_1, φ_2 - phase mismatch.

If there is no mismatch ($\alpha_1 = \alpha_2 = 1, \varphi_1 = \varphi_2 = 0$), the input signal is multiplied by $\sin^2(\omega_L t) + \cos^2(\omega_L t)$ which equals to 1, thus the signal is not modified.

Let us assume a small mismatch, that is:

$$\alpha_1 \cong \alpha_2 \cong 1 \quad \text{Eq. I.2}$$

$$\varphi_1 \cong \varphi_2 \ll 1 \quad \text{Eq. I.3}$$

Using basic trigonometric identities and Taylor series approximation:

$$F_0(t) = \left\{ \sin^2(\omega_L t) + \frac{1}{2} \alpha_1 \alpha_2 \{ \cos[2\omega_L t + (\varphi_1 + \varphi_2)] + \cos(\varphi_1 - \varphi_2) \} \right\} F_{in}(t) \quad \text{Eq. I.4}$$

$$\cos x \approx 1 - \frac{1}{2} x^2 \quad \text{Eq. I.5}$$

$$F_0(t) = \left\{ \sin^2(\omega_L t) + \frac{1}{2} \alpha_1 \alpha_2 \{1 + \cos[2\omega_L t + (\varphi_1 + \varphi_2)]\} - \frac{1}{4} \alpha_1 \alpha_2 (\varphi_1 - \varphi_2)^2 \right\} F_{in}(t) \quad \text{Eq. I.6}$$

$$\cos^2 x = \frac{1}{2} (\cos 2x + 1) \quad \text{Eq. I.7}$$

$$\alpha_1 \alpha_2 (\varphi_1 - \varphi_2)^2 \cong 0 \quad \text{Eq. I.8}$$

$$F_0(t) = \left\{ \sin^2(\omega_L t) + \alpha_1 \alpha_2 \cos^2[\omega_L t + \frac{1}{2}(\varphi_1 + \varphi_2)] \right\} F_{in}(t) \quad \text{Eq. I.9}$$

$$\begin{aligned} \cos^2(x + \varepsilon) &= \frac{1}{2} [1 + \cos 2(x + \varepsilon)] = \frac{1}{2} [1 + \cos 2x \cos 2\varepsilon - \sin 2x \sin 2\varepsilon] \\ &\cong \frac{1}{2} [1 + \cos 2x (1 - 2\varepsilon^2) - 2\varepsilon \sin 2x] \\ &= \frac{1}{2} (1 + \cos 2x) - \varepsilon^2 \cos 2x - \varepsilon \sin 2x \end{aligned} \quad \text{Eq. I.10}$$

$$\varepsilon^2 \cos 2x \cong 0 \quad \text{Eq. I.11}$$

$$\cos^2(x + \varepsilon) \cong \cos^2 x - \varepsilon \sin 2x \quad \text{Eq. I.12}$$

$$\alpha_1 = 1 + \varepsilon_1; \varepsilon_1 \approx 0 \quad \text{Eq. I.13}$$

$$\alpha_2 = 1 + \varepsilon_2; \varepsilon_2 \approx 0 \quad \text{Eq. I.14}$$

$$F_0(t) = \left[\sin^2(\omega_L t) + (1 + \varepsilon_1)(1 + \varepsilon_2) [\cos^2 \omega_L t - \frac{1}{2}(\varphi_1 + \varphi_2) \sin 2\omega_L t] \right] F_{in}(t) \quad \text{Eq. I.15}$$

$$(1 + \varepsilon_1)(1 + \varepsilon_2) \cong 1 + \varepsilon_1 + \varepsilon_2 \quad \text{Eq. I.16}$$

$$\begin{aligned} F_0(t) &= \left[\sin^2(\omega_L t) + \cos^2 \omega_L t - \frac{1}{2}(\varphi_1 + \varphi_2) \sin 2\omega_L t + (\varepsilon_1 + \varepsilon_2) \cos^2 \omega_L t \right. \\ &\quad \left. - \frac{1}{2}(\varepsilon_1 + \varepsilon_2)(\varphi_1 + \varphi_2) \sin 2\omega_L t \right] F_{in}(t) \end{aligned} \quad \text{Eq. I.17}$$

$$(\varepsilon_1 + \varepsilon_2)(\varphi_1 + \varphi_2) \cong 0 \quad \text{Eq. I.18}$$

$$F_0(t) = \left[1 - \frac{1}{2}(\varphi_1 + \varphi_2) \sin 2\omega_L t + (\varepsilon_1 + \varepsilon_2) \cos^2 \omega_L t \right] F_{in}(t) \quad \text{Eq. I.19}$$

$$\cos^2 x = \frac{1}{2} (\cos 2x + 1) \quad \text{Eq. I.20}$$

$$F_0(t) = \left\{ 1 + \frac{1}{2}(\varepsilon_1 + \varepsilon_2)[\cos 2\omega_L t + 1] - \frac{1}{2}(\varphi_1 + \varphi_2)\sin 2\omega_L t \right\} F_{in}(t) \quad \text{Eq. I.21}$$

$$F_0(t) = \left\{ 1 + \frac{1}{2}[(\varepsilon_1 + \varepsilon_2) + (\varepsilon_1 + \varepsilon_2)\cos 2\omega_L t - (\varphi_1 + \varphi_2)\sin 2\omega_L t] \right\} F_{in}(t) \quad \text{Eq. I.22}$$

All the terms in brackets, despite the first one, describe an error signal $F_E(t)$.

$$F_E(t) = \frac{1}{2}[(\varepsilon_1 + \varepsilon_2) + (\varepsilon_1 + \varepsilon_2)\cos 2\omega_L t - (\varphi_1 + \varphi_2)\sin 2\omega_L t] F_{in}(t) \quad \text{Eq. I.23}$$

Let us transform the Eq. 3.9 to find a spectrum of an error signal.

$$\cos 2x = \frac{e^{j2x} + e^{-j2x}}{2} \quad \text{Eq. I.24}$$

$$\sin 2x = \frac{e^{j2x} - e^{-j2x}}{2j} \quad \text{Eq. I.25}$$

$$\begin{aligned} F_E(t) &= \frac{1}{2}[(\varepsilon_1 + \varepsilon_2) + (\varepsilon_1 + \varepsilon_2)\cos 2\omega_L t - (\varphi_1 + \varphi_2)\sin 2\omega_L t] \\ &= \frac{1}{2} \left\{ (\varepsilon_1 + \varepsilon_2) \right. \\ &\quad \left. + \frac{1}{2}[(\varepsilon_1 + \varepsilon_2)(e^{j2\omega_L t} + e^{-j2\omega_L t}) \right. \\ &\quad \left. - (\varphi_1 + \varphi_2)\frac{(e^{j2\omega_L t} - e^{-j2\omega_L t})}{j}] \right\} \end{aligned} \quad \text{Eq. I.26}$$

$$\begin{aligned} F_E(t) &= \frac{1}{2} \left[(\varepsilon_1 + \varepsilon_2)(e^{j2\omega_L t} + e^{-j2\omega_L t}) - (\varphi_1 + \varphi_2)\frac{(e^{j2\omega_L t} - e^{-j2\omega_L t})}{j} \right] \\ &= \frac{1}{2} \{ e^{j2\omega_L t} [(\varepsilon_1 + \varepsilon_2) + j(\varphi_1 + \varphi_2)] \\ &\quad + e^{-j2\omega_L t} [(\varepsilon_1 + \varepsilon_2) - j(\varphi_1 + \varphi_2)] \} \end{aligned} \quad \text{Eq. I.27}$$

And finally:

$$\begin{aligned} F_E(t) &= \left\{ \frac{1}{2}(\varepsilon_1 + \varepsilon_2) + \frac{1}{4}[(\varepsilon_1 + \varepsilon_2) + j(\varphi_1 + \varphi_2)]e^{j2\omega_L t} \right. \\ &\quad \left. + \frac{1}{4}[(\varepsilon_1 + \varepsilon_2) - j(\varphi_1 + \varphi_2)]e^{-j2\omega_L t} \right\} F_{in}(t) \end{aligned} \quad \text{Eq. I.28}$$

Appendix II

ECL termination techniques

ECL is an acronym for Emitter Coupled Logic. Its main advantage comparing to TTL and CMOS technologies is much higher switching speed [32]. This is achieved with avoiding saturation effects by limiting the emitter current. Proper termination and current biasing of ECL driver outputs is needed for the good static and dynamic performance. A standard output driver consists of the emitter follower which must remain in the active region. It requires an external resistive path from the output to a voltage more negative than the lowest low logic level. This chapter describes termination techniques used in the project.

II.1 Parallel termination.

Parallel termination offers the best circuit performance. In this technique a driver is connected to a receiver via a transmission line terminated to a V_{TT} termination supply by a resistor R_t of a value equal to the line characteristic impedance.

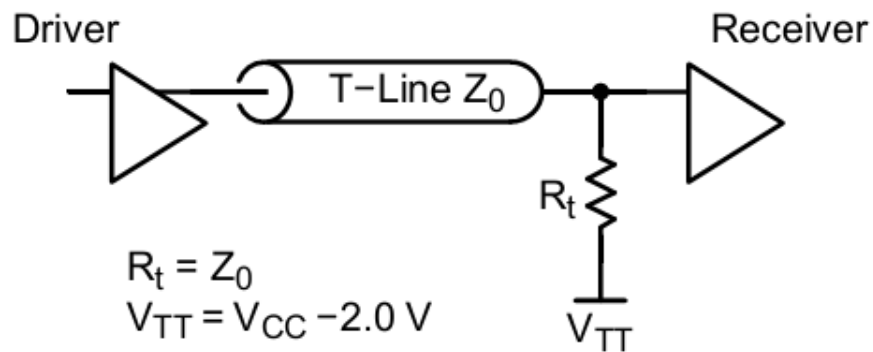


Fig. II.1 Parallel termination [33].

The driver output biasing current flows through the line. The V_{TT} supply must be more negative than the lowest output voltage and always sink current in order to keep the output transistor in the active region. Typically V_{TT} is 2.0 V below the V_{CC} supply.

A usage of an additional V_{TT} voltage supply can be omitted by using a resistive divider. Connected to the V_{CC} it can deliver required biasing voltage and match the line characteristic impedance, as shown in Fig. II.2.

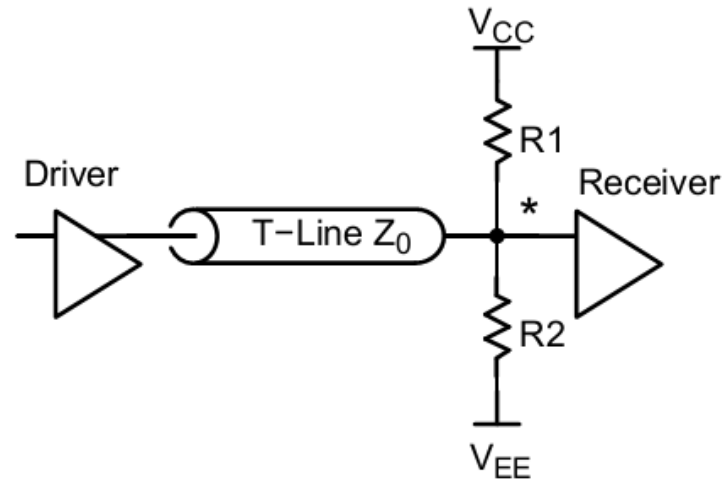


Fig. II.2 Thevenin equivalent parallel termination [33].

The resistance of the parallel connection of R_1 and R_2 must be equal to the line characteristic impedance and a DC voltage at R_2 equals to V_{TT} . Such configuration does not require additional power supply, but it consumes more energy. However, the additional power is fully dissipated in external resistors R_1 and R_2 , thus it does not affect the IC parameters.

II.2 Series termination.

Series termination techniques can be used for driving lines with impedance discontinuities and long interconnections. Possibility of driving a couple of lines with a single driver and minimising crosstalk problems are among the advantages of the techniques. In a series termination technique a termination resistor is placed between the driver and the transmission line. Transmission line is directly connected to the receiver.

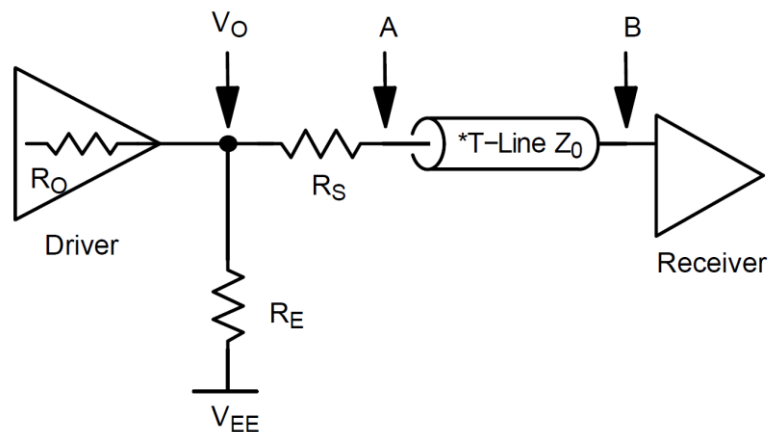


Fig. II.3 Series termination [33].

A driver output level transition creates the voltage change. This results in a change of the voltage on the input of transmission line (point A) which propagates through the transmission line to the receiver input (point B).

$$\Delta V_A = \Delta V_O * \left(\frac{Z_0}{R_O + R_S + Z_0} \right) \quad \text{Eq. II.1}$$

ΔV_A – voltage change on the transmission line input;

ΔV_O – voltage change on the driver output;

Z_0 – line characteristic impedance;

R_O – driver output impedance;

R_S – termination resistor.

The sum of the termination resistor R_S and the driver output impedance R_O is equal to the characteristic impedance of the transmission line Z_0 . Substitution $Z_0 = R_S + R_O$ into the previous equation yields:

$$\Delta V_A = \frac{\Delta V_O}{2} \quad \text{Eq. II.2}$$

An amplitude of an incident wave is a half the driver output voltage change. High input impedance of the receiver results in a full positive reflection, doubling the voltage at the end of the transmission line. The voltage change at the point B (ΔV_B) is approximately equal to the ΔV_O . Because of the fact that the sum of the driver output impedance and series termination resistor equals to the line characteristic impedance, the reflected wave energy is dissipated and there are no further reflections.

The function of the current bias resistor R_E is to establish logic high (V_{OH}) and low levels (V_{OL}). Voltage change developed on R_E during the transition is propagated through a series resistor and the transmission line to the receiver input. When the driver output switches to the low state, the negative voltage transition is current limited by R_E , R_S and Z_0 . Emitter pull-down resistor has two goals. Firstly it must maximize the negative voltage transition. The second goal is to prevent the output transistor from entering the cut-off operating region in a low state.

When the driver output emitter follower enters into cut-off it can be considered open and the line characteristic impedance behaves as a linear resistor returned to V_{OH} . This situation is shown in Fig. II.4.

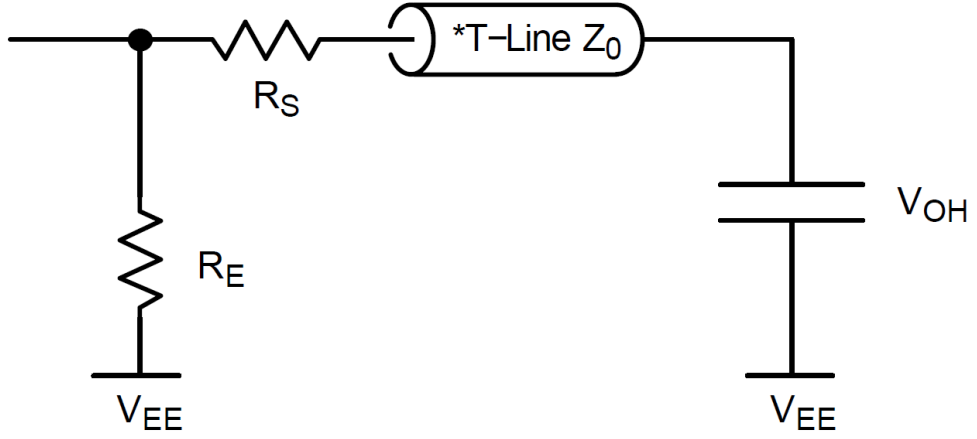


Fig. II.4 Equivalent circuit of emitter follower entered into cut-off [33].

There is a maximum current in the moment of opening the switch. This current must be able to generate a transient voltage equal to half of the logic swing – the voltage will double at the receiver due to the full positive reflection. To compensate the reflections caused by discontinuities and load capacitances the transient voltage should be increased. Author of the AND8020/D application note suggests increasing the voltage by a factor of 25%. Thus the required initial current I_{init} is given by:

$$I_{init} = \frac{1.25 * \frac{V_{swing}}{2}}{Z_0} \quad \text{Eq. II.3}$$

The maximum current occurs during a negative going transition and is given by:

$$I_{max} = \frac{(V_{OH} - V_{EE})}{(R_E + R_S + Z_0)} \quad \text{Eq. II.4}$$

Current flowing in the circuit at this moment must be higher than the initial current:

$$I_{max} > I_{init} \quad \text{Eq. II.5}$$

$$\frac{(V_{OH} - V_{EE})}{(R_E + R_S + Z_0)} > \frac{1.25 * \frac{V_{swing}}{2}}{Z_0} \quad \text{Eq. II.6}$$

A value of R_E can be determined from above inequality:

$$R_E < \left(\frac{2 * (V_{OH} - V_{EE})}{1.25 * V_{swing}} - 1 \right) * Z_0 - R_S \quad \text{Eq. II.7}$$

II.3 Parallel fan-out using series termination.

The series termination technique can be extended to drive a couple of lines without using buffers. Figure 3 shows this situation.

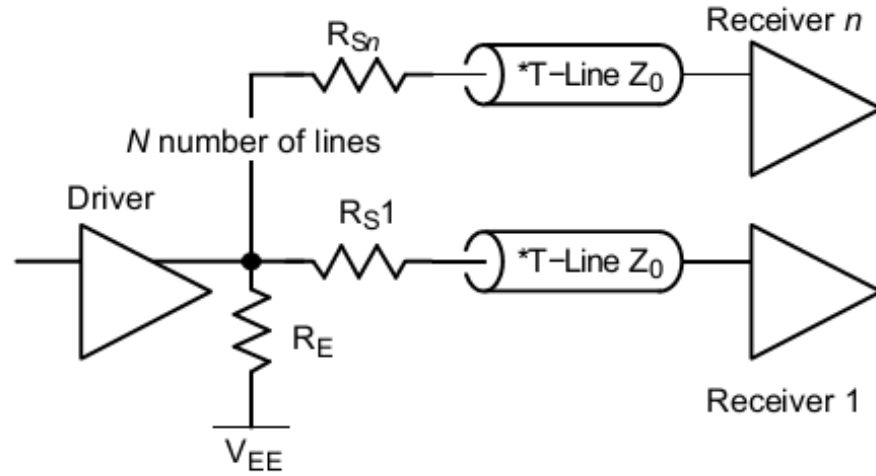


Fig. II.5 Parallel fan-out using series termination [33].

A single ECL gate is used to drive several transmission lines in parallel fan-out. The current in the output emitter follower must be kept below the maximum value. The R_E value can be calculated as a parallel connection of emitter pull-down resistors calculated separately for each of the lines according to Eq. II.7. When there are n lines of identical characteristic impedance, the R_E is calculated as:

$$R_E < \frac{\left(\frac{2 \cdot (V_{OH} - V_{EE})}{1.25 \cdot V_{swing}} - 1 \right) \cdot Z_0 - R_S}{n} \quad \text{Eq. II.8}$$

Appendix III

Performance characteristics

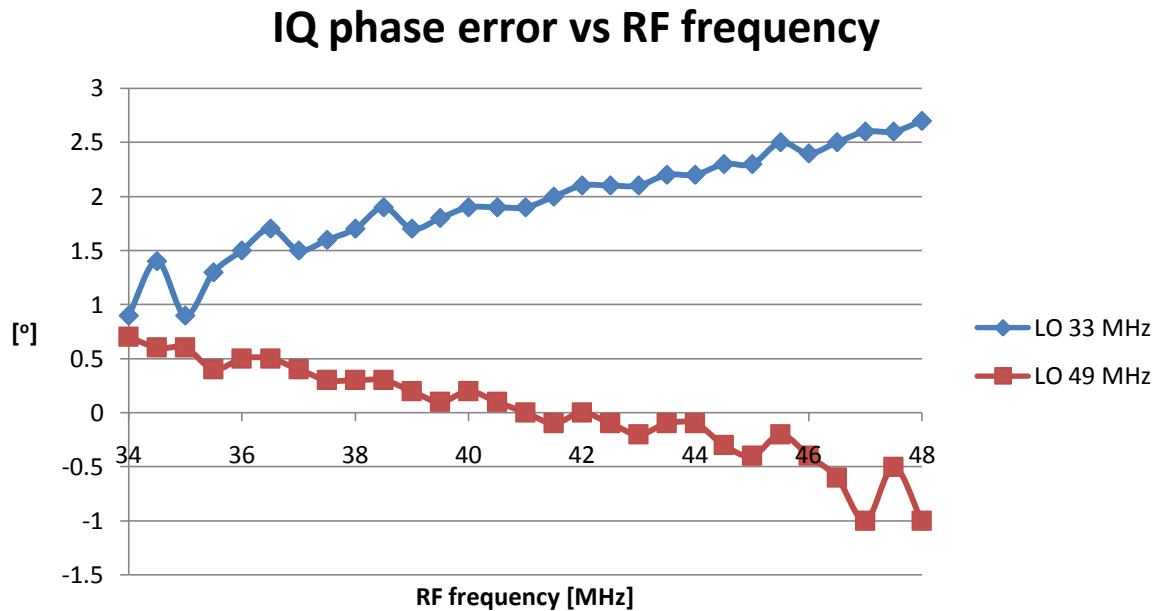


Fig. III.1 Demodulator phase error.

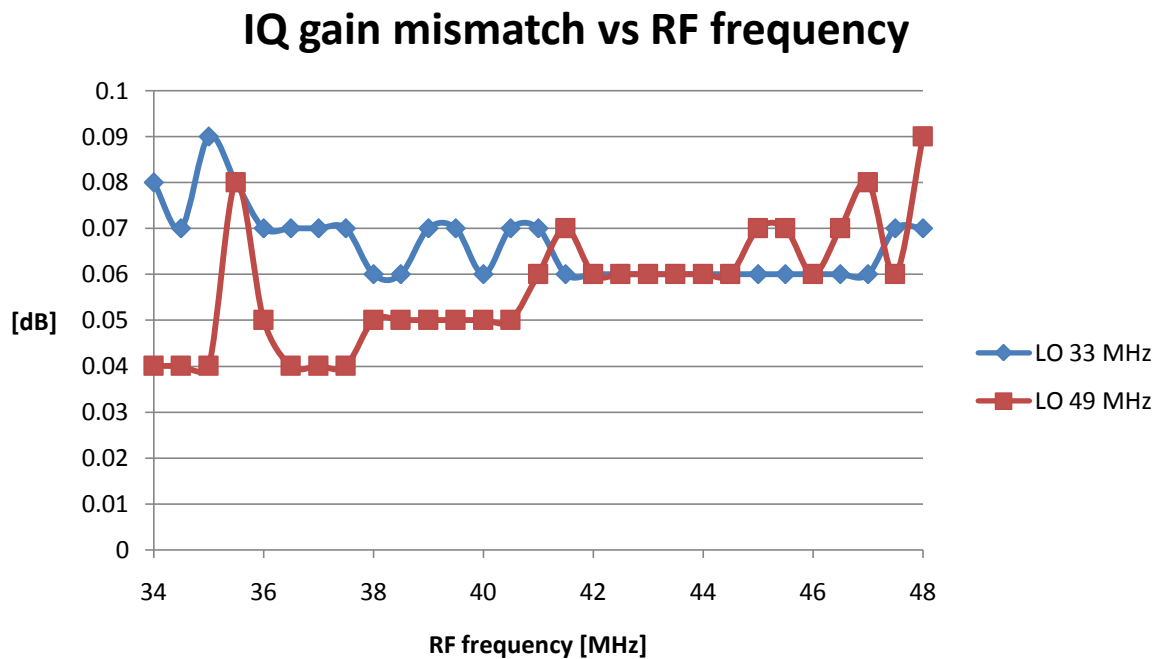


Fig. III.2 Demodulator amplitude mismatch.

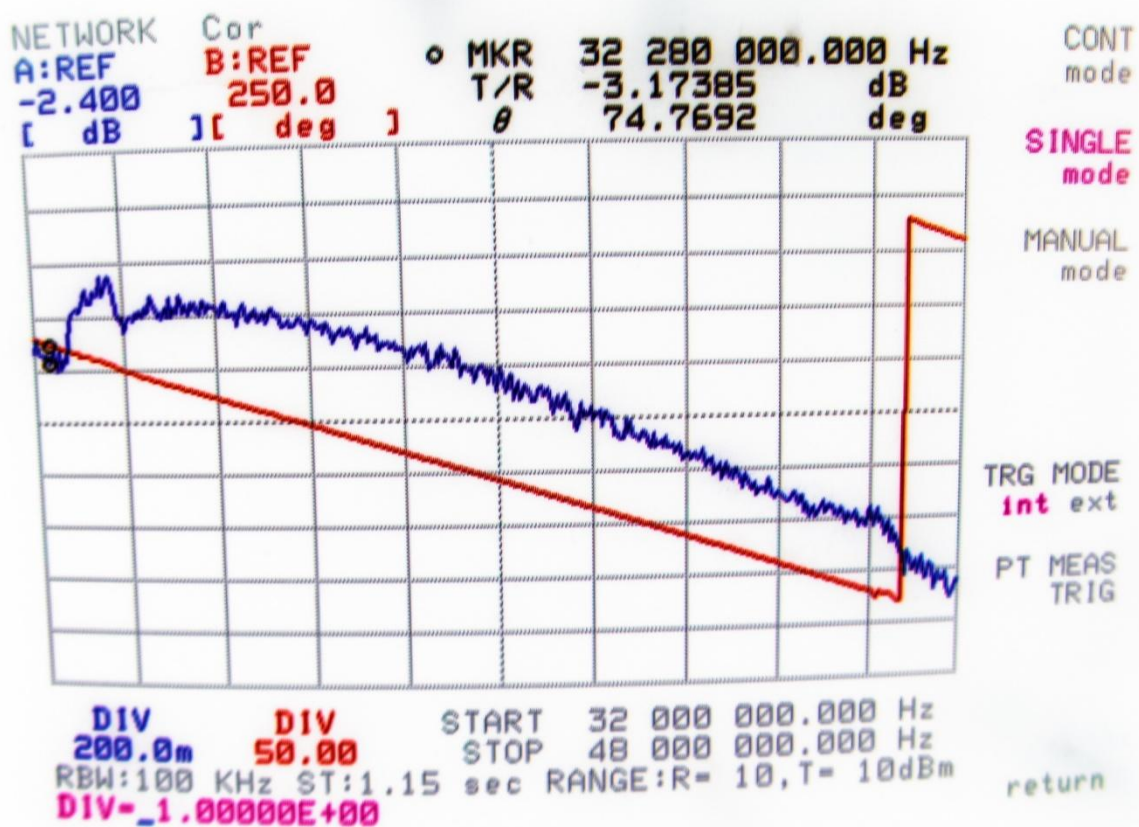


Fig. III.3 Transfer function, LO frequency 33 MHz (blue trace – gain, red – phase).

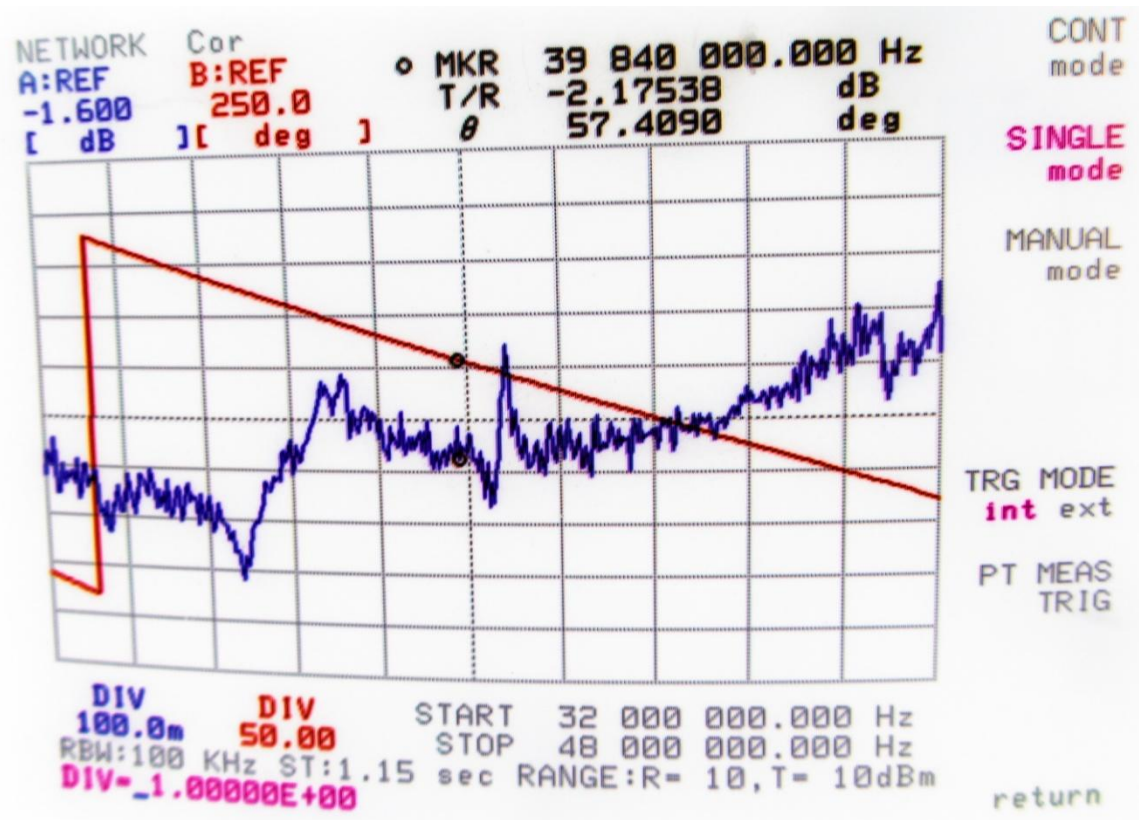


Fig. III.4 Transfer function, LO frequency 49 MHz (blue trace – gain, red – phase).

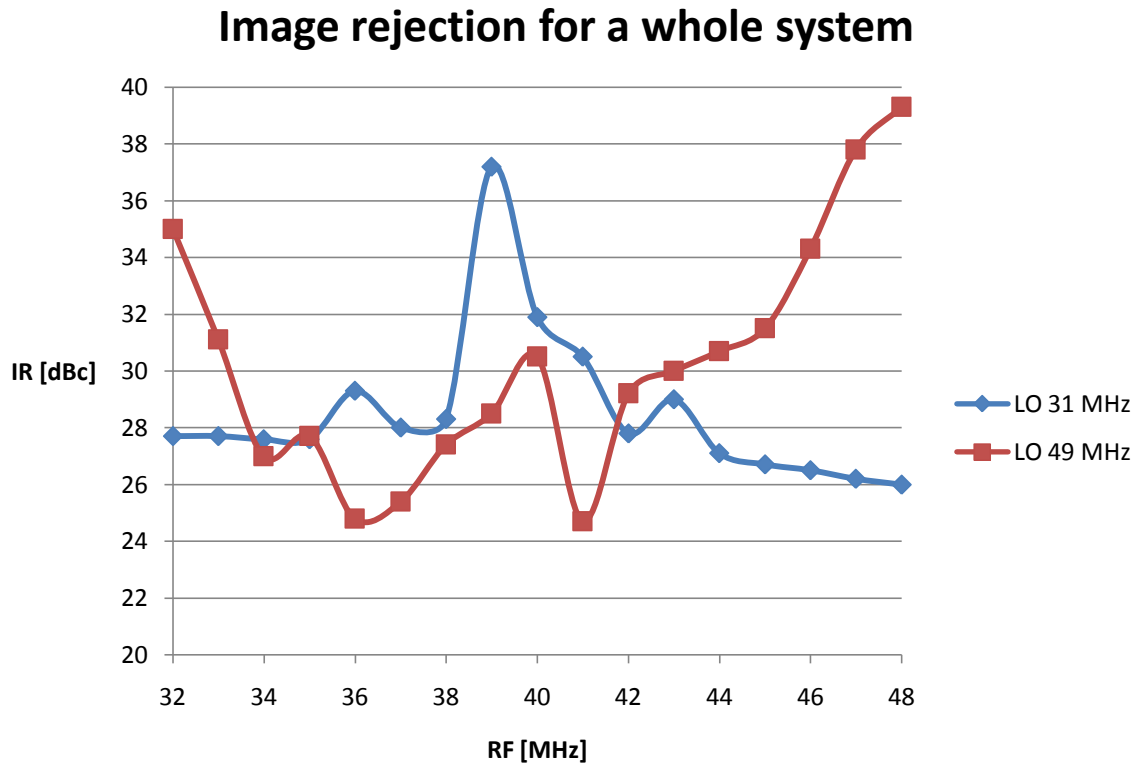


Fig. III.5 Image rejection vs. RF frequency for a whole system.

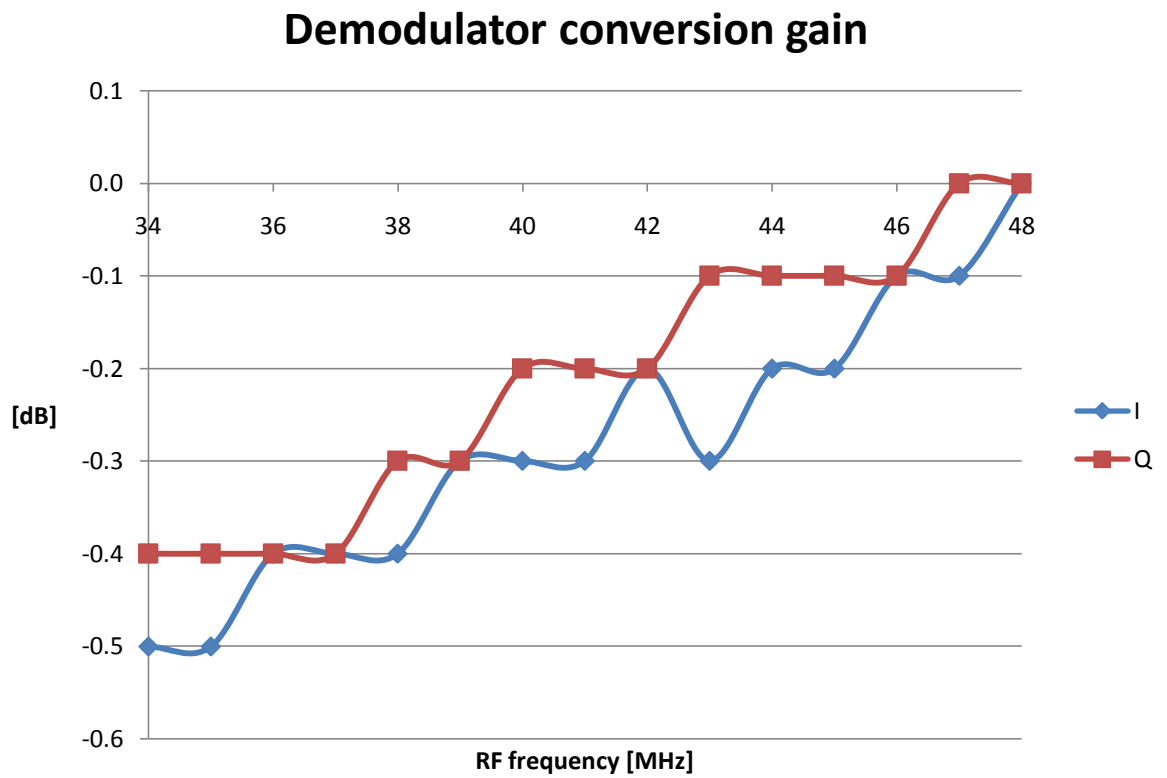


Fig. III.6 Demodulator conversion gain.

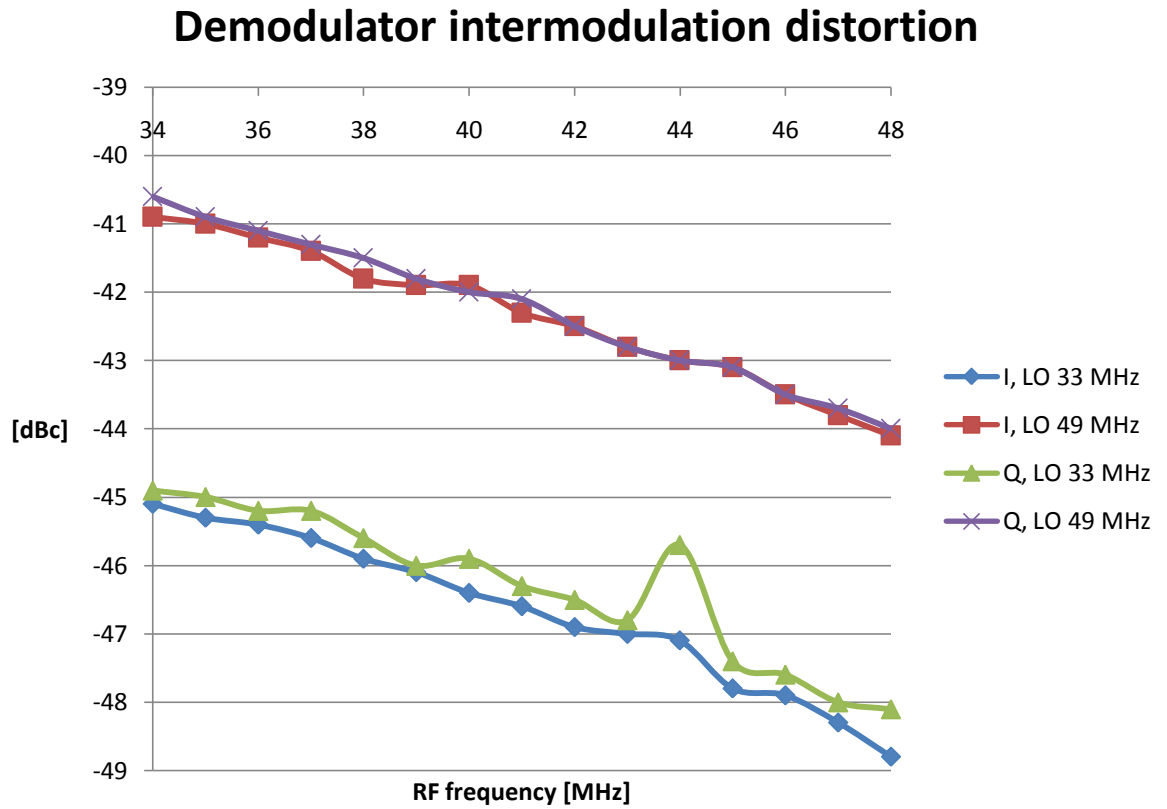


Fig. III.7 Demodulator intermodulation distortion.

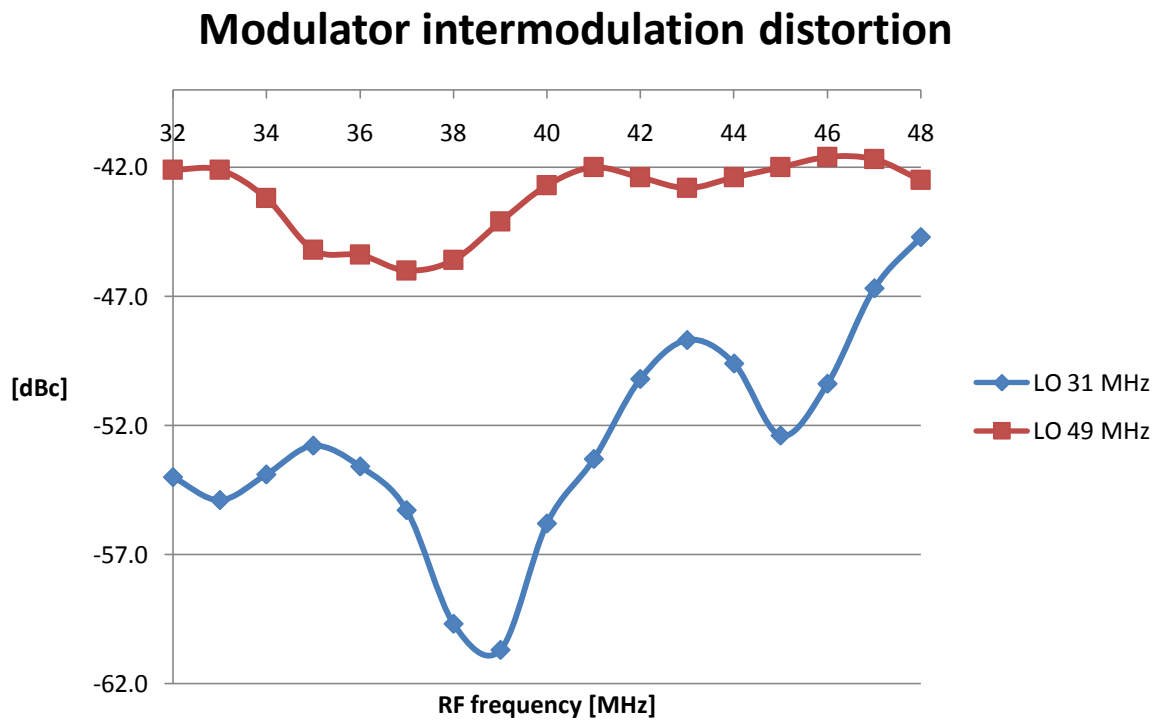


Fig. III.8 Modulator intermodulation distortion.

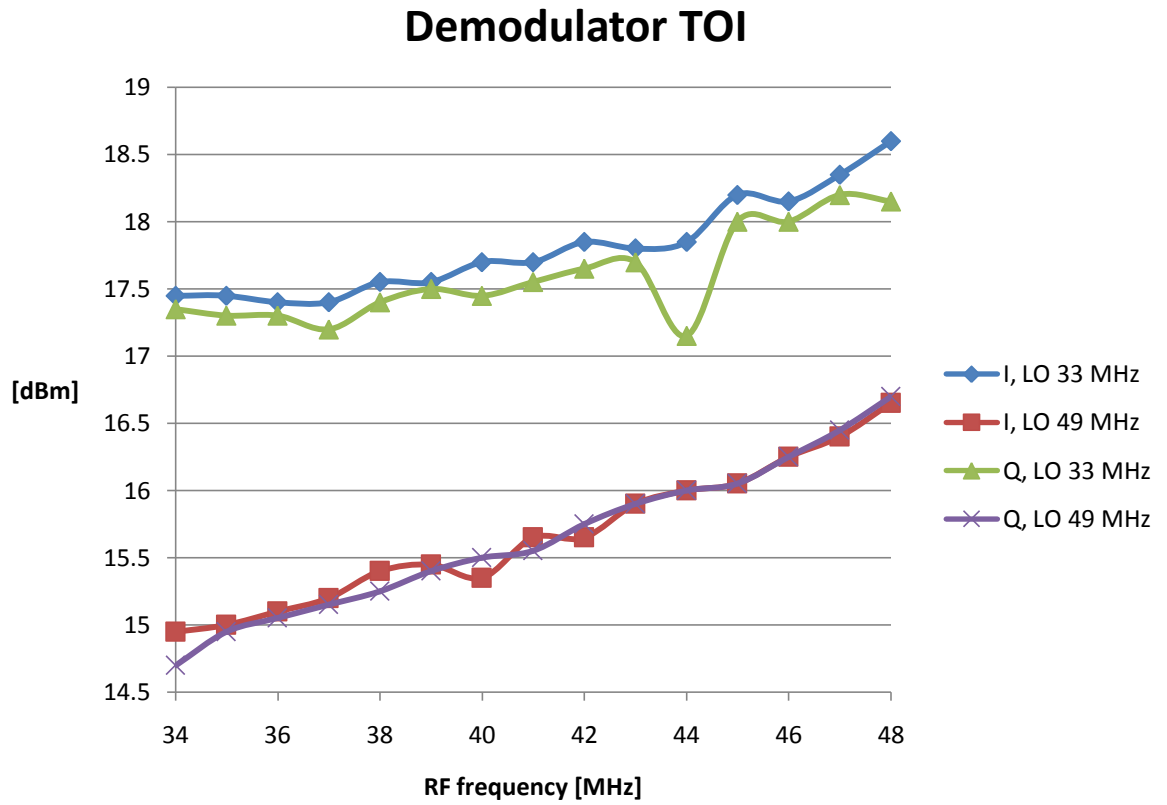


Fig. III.9 Demodulator TOI.

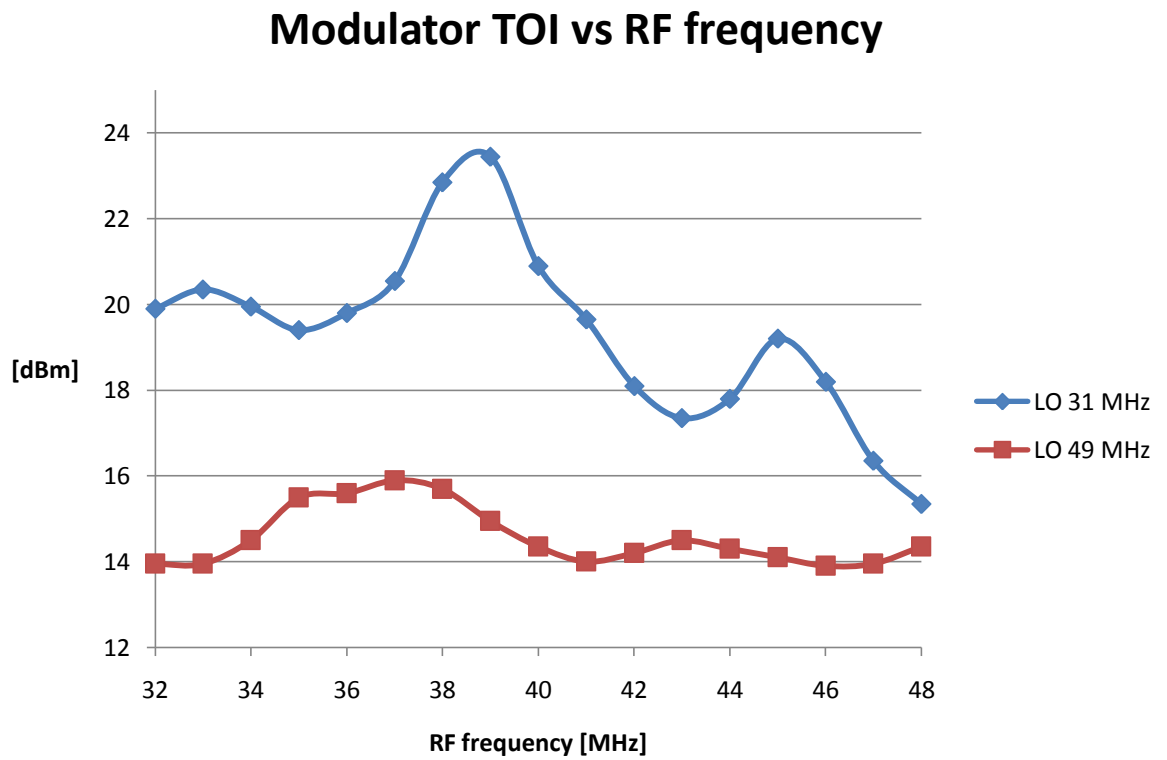


Fig. III.10 Modulator TOI.

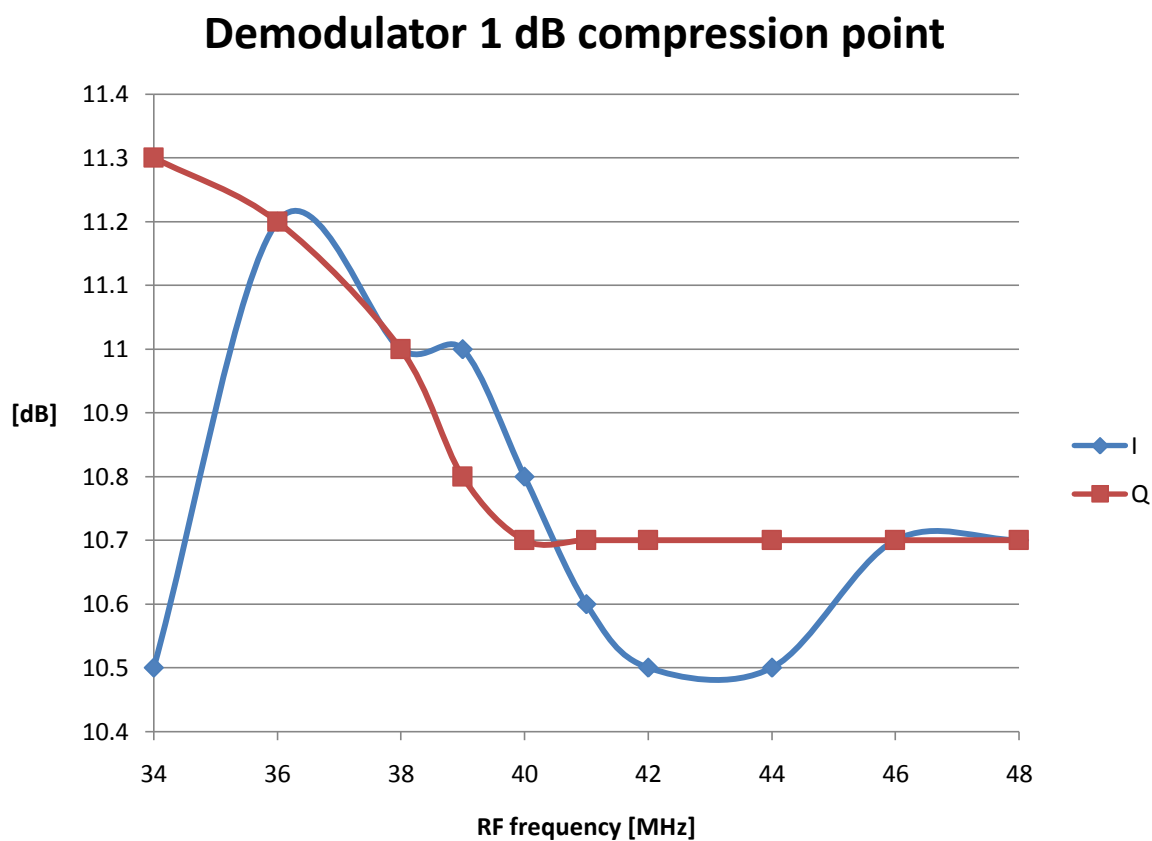


Fig. III.11 Demodulator 1 dB compression point.

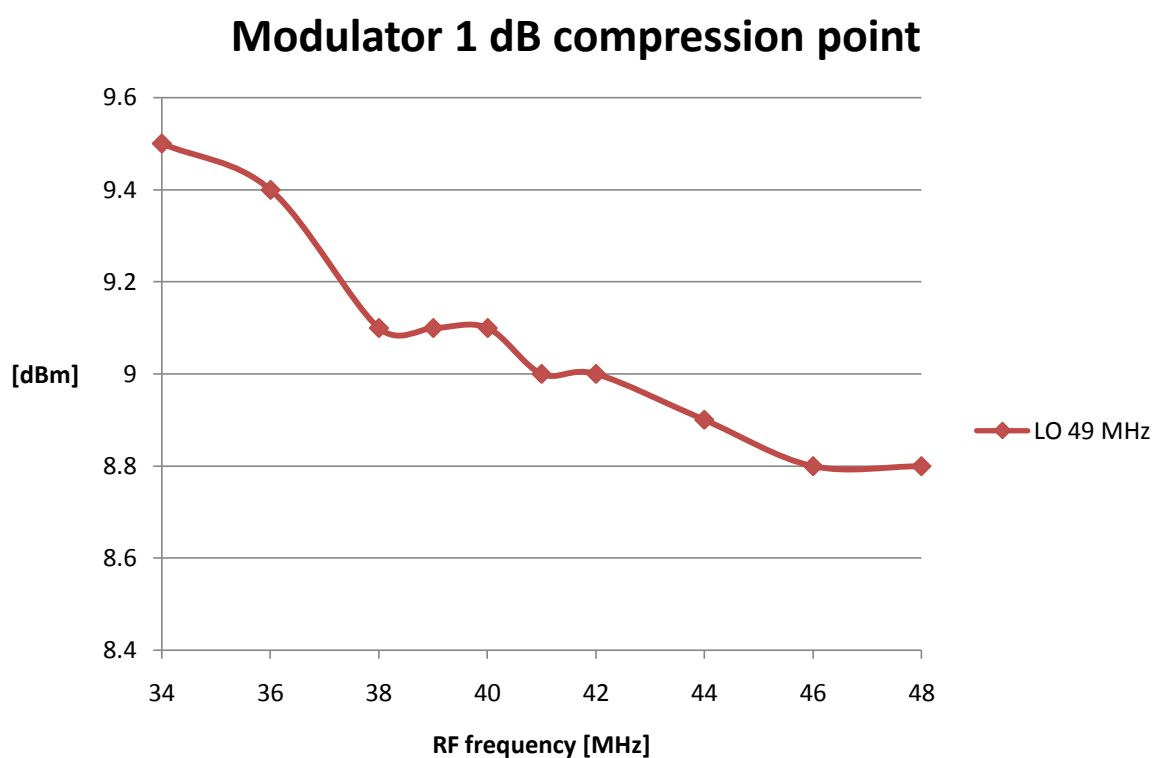


Fig. III.12 Modulator 1 dB compression point.

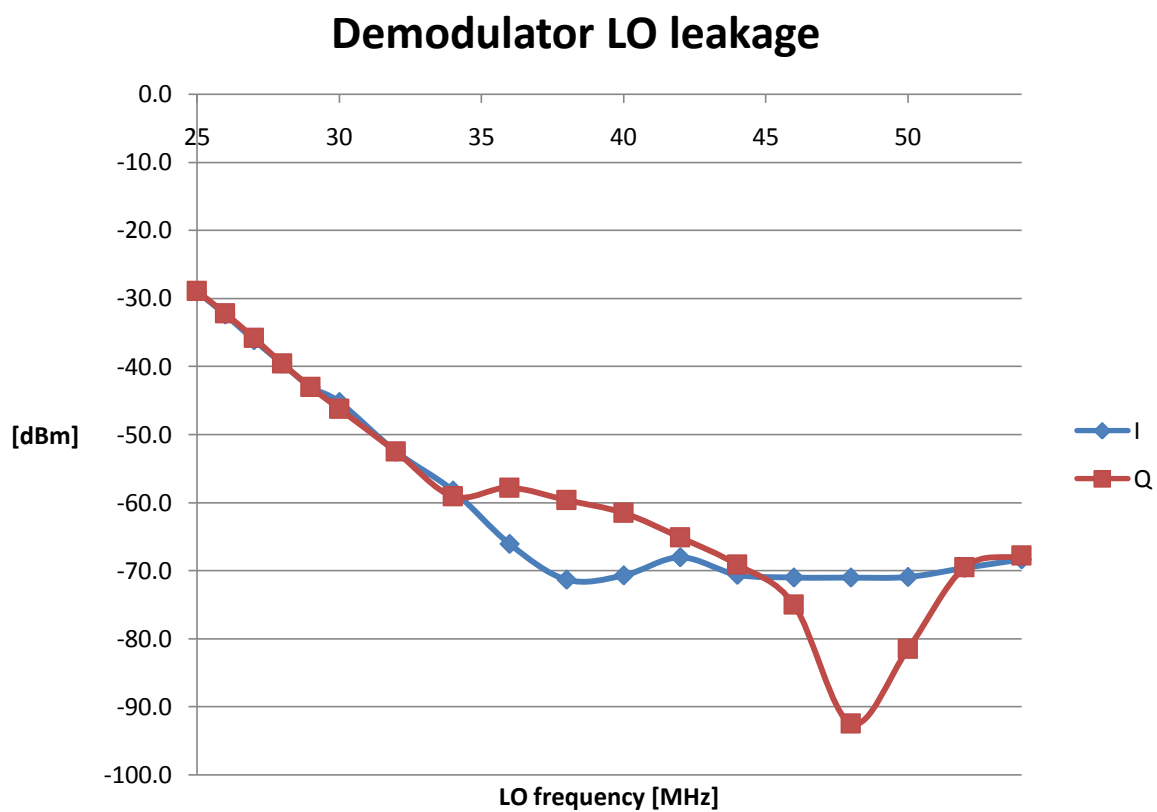


Fig. III.13 LO to IF leakage.

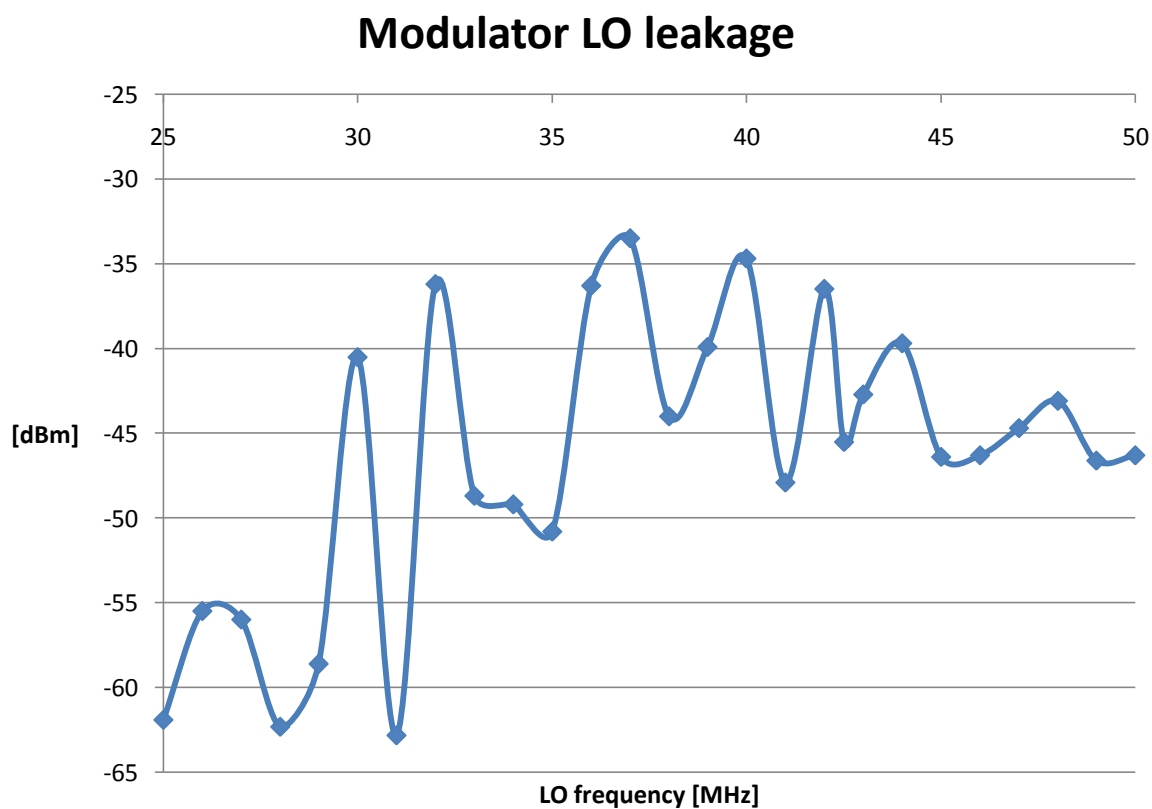


Fig. III.14 LO to RF output leakage.

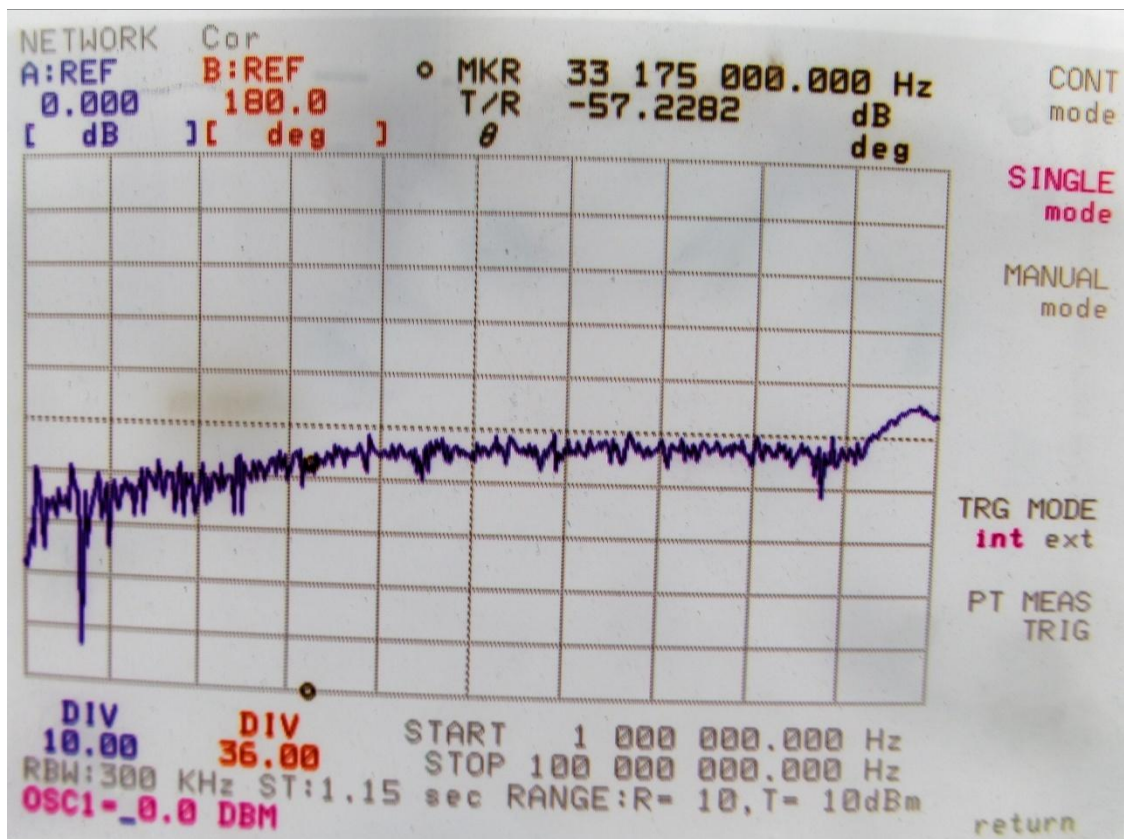


Fig. III.15 I output to RF input isolation.

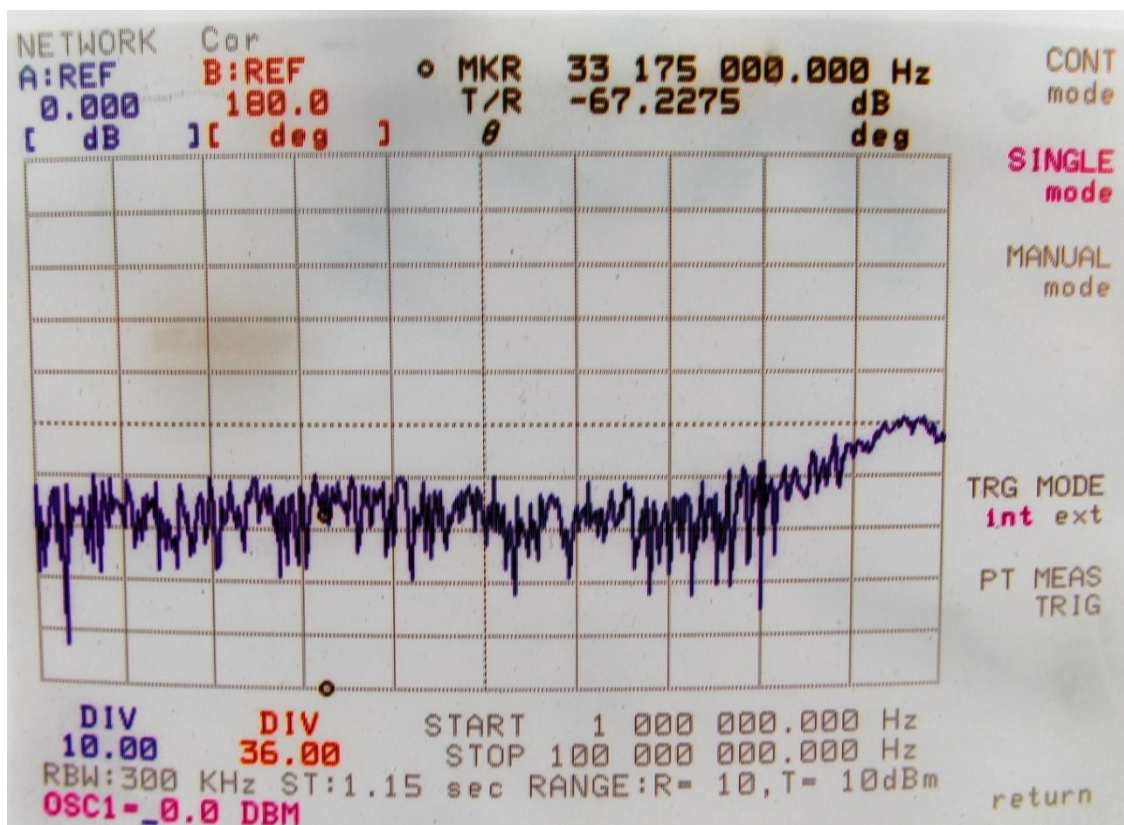


Fig. III.16 Q output to RF input isolation.

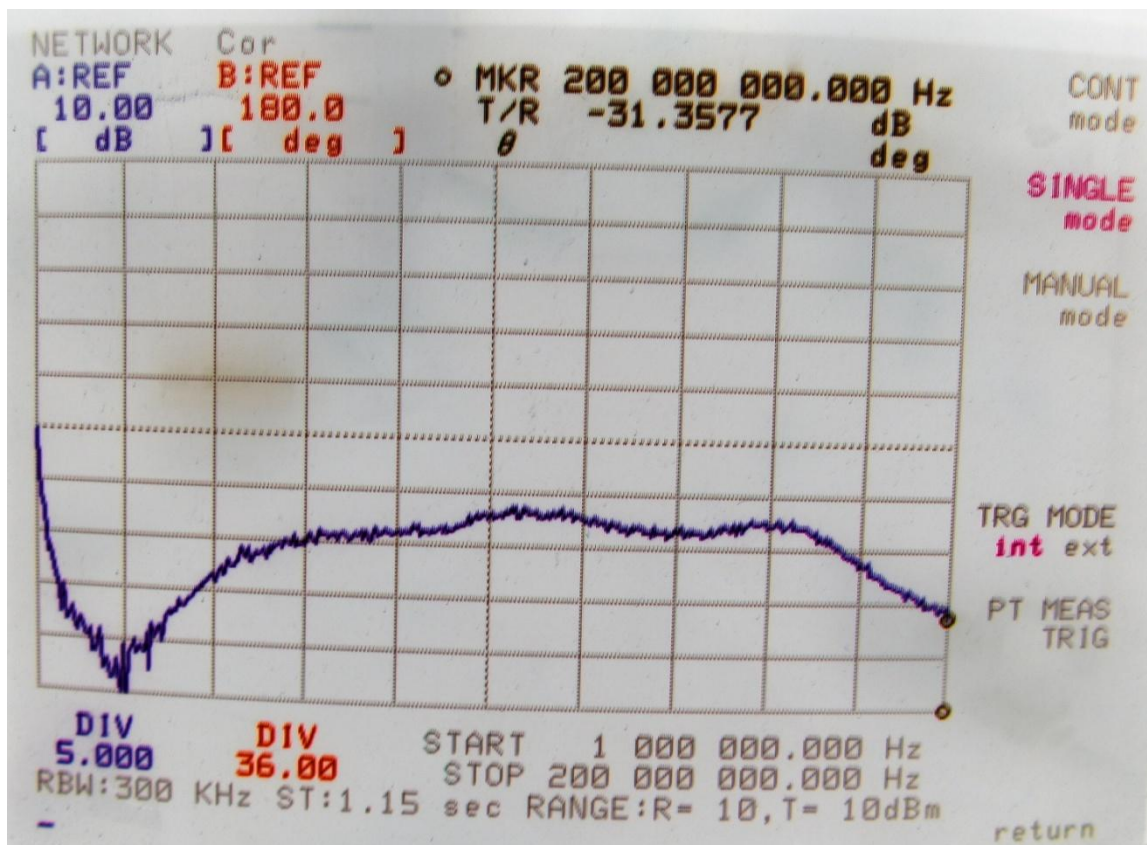


Fig. III.17 2LO input return loss.

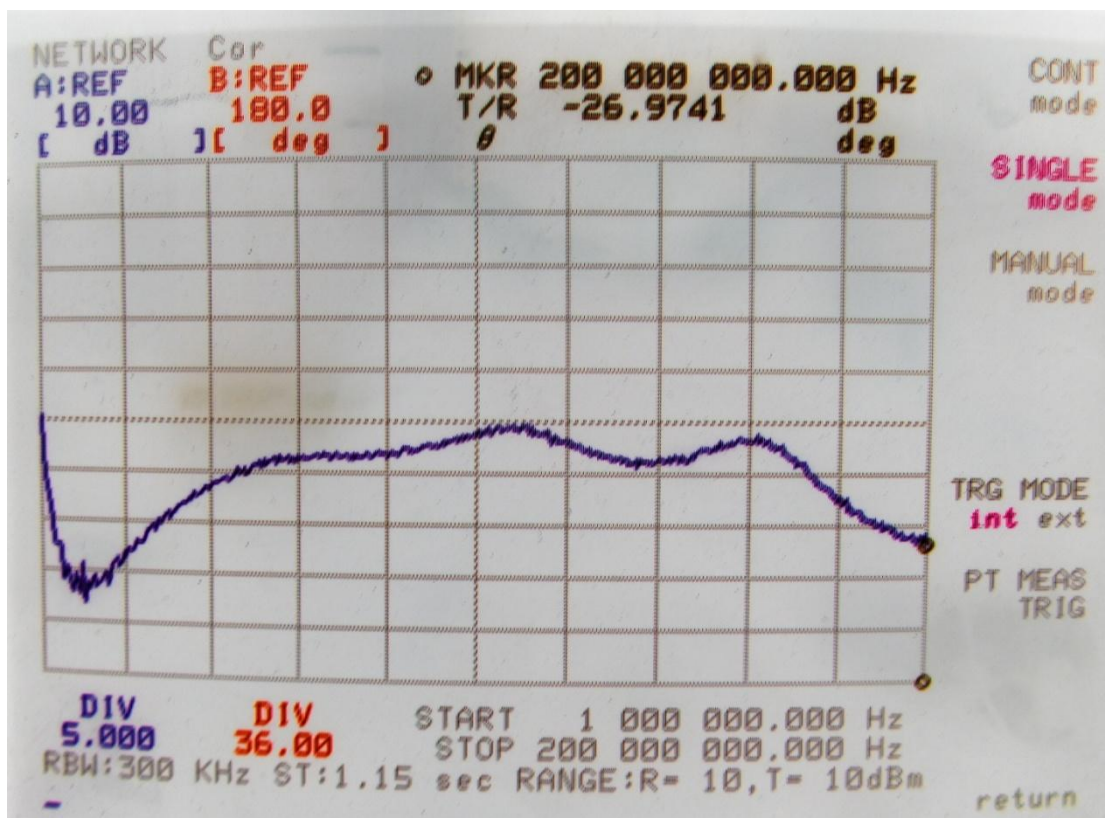


Fig. III.18 RF input return loss.

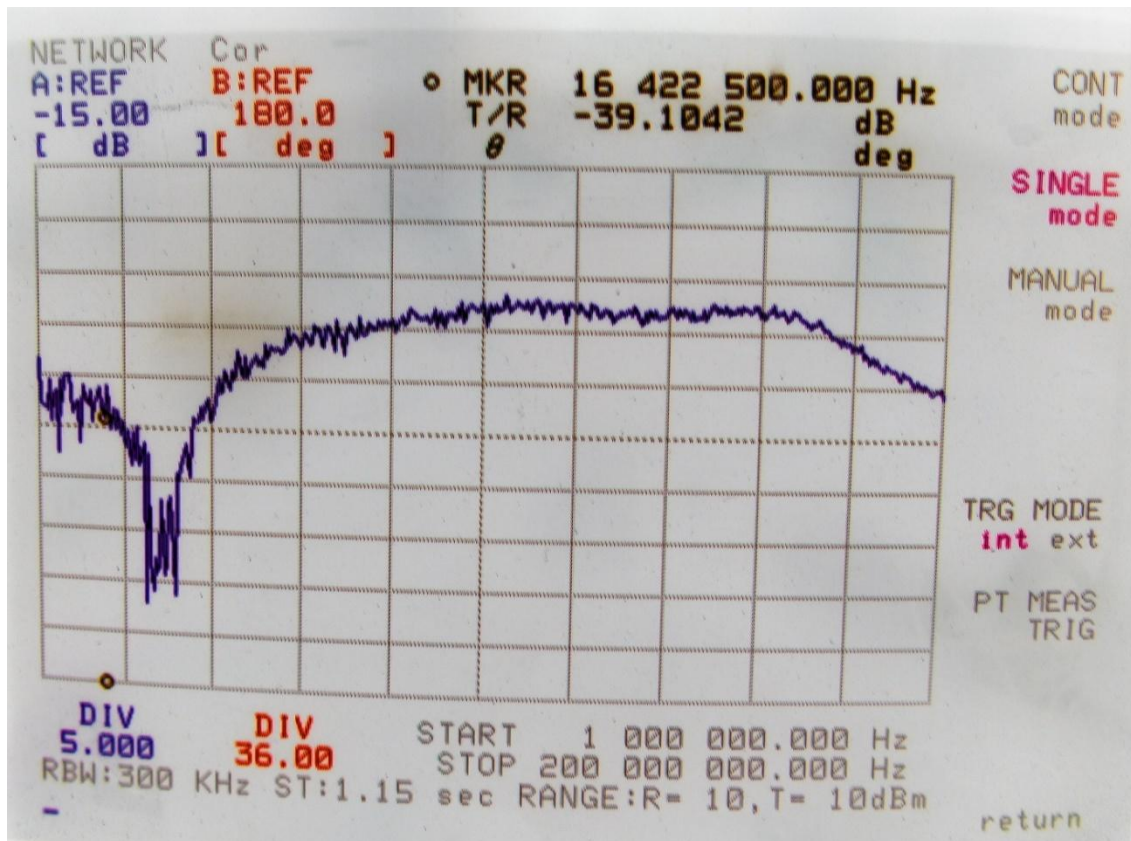


Fig. III.19 I input return loss.

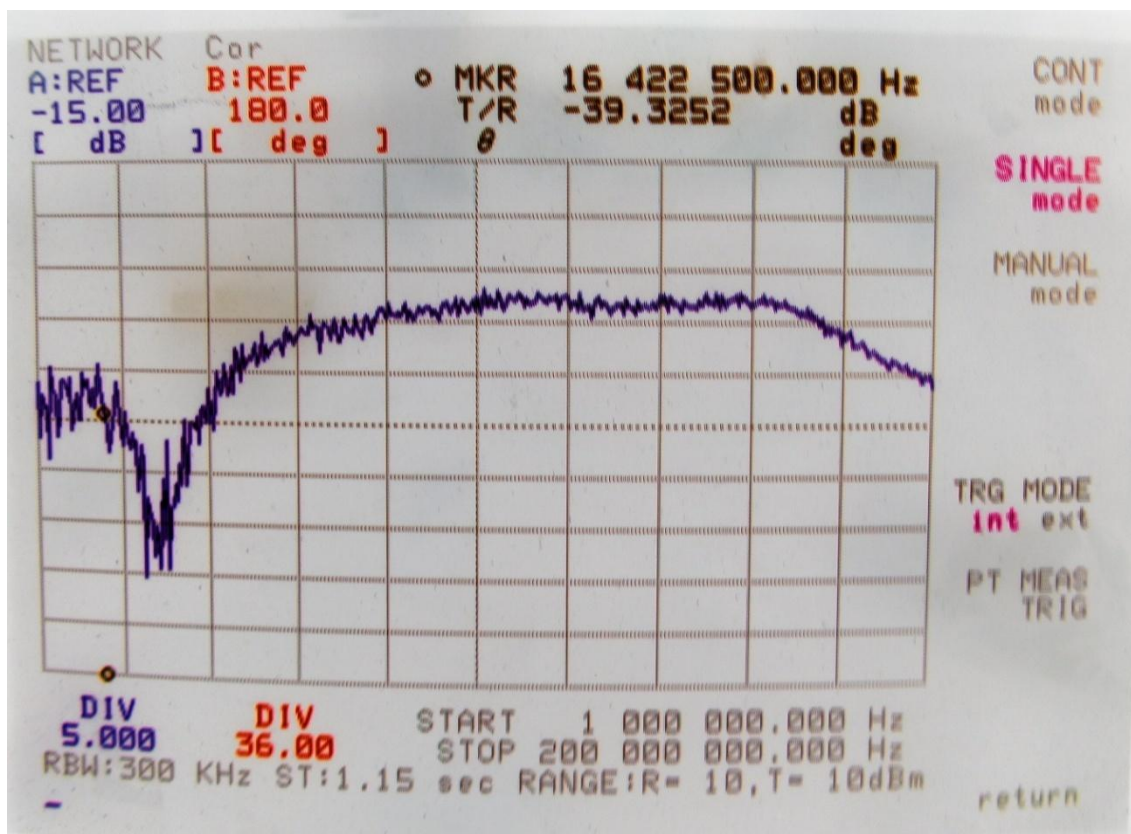
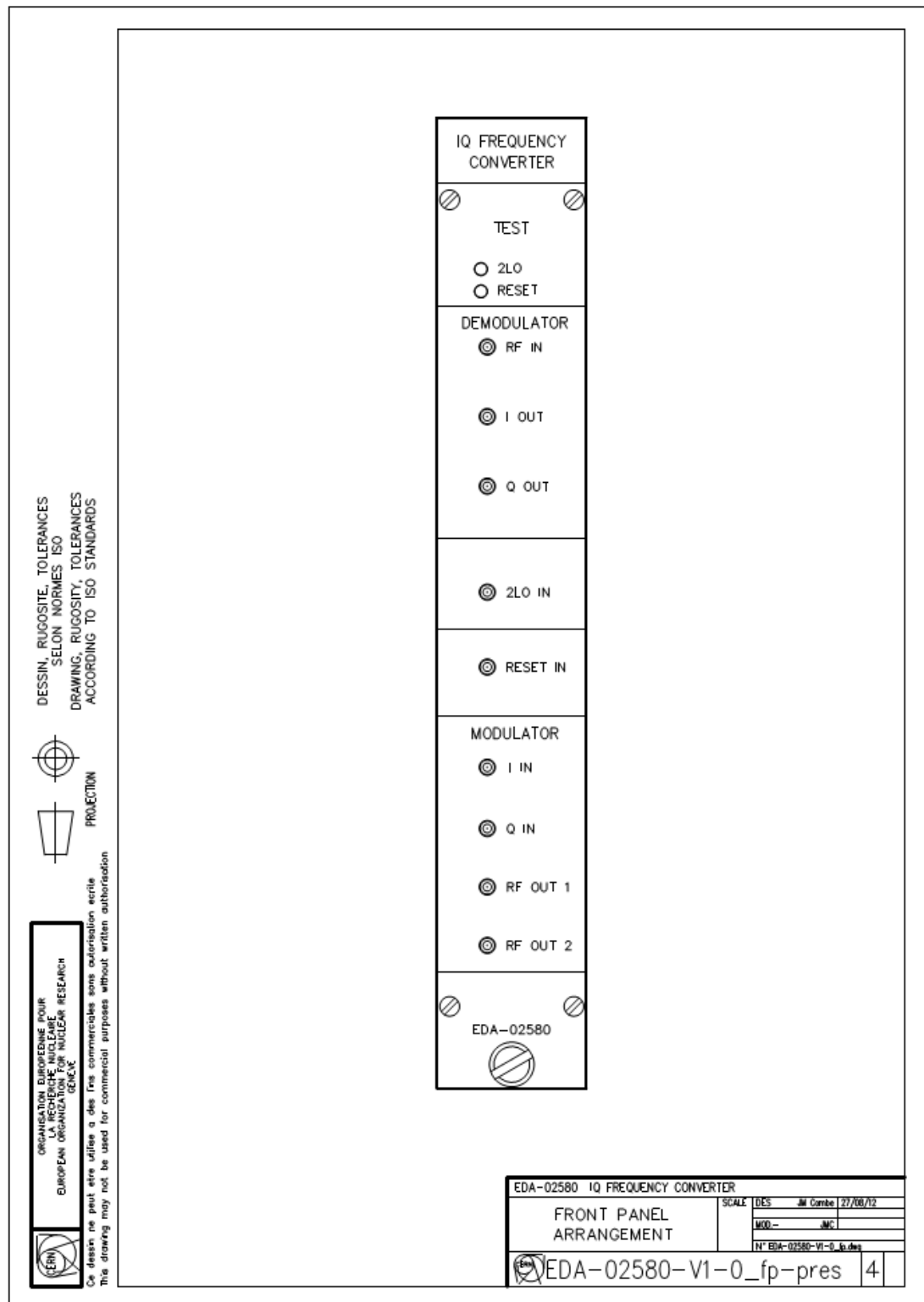


Fig. III.20 Q input return loss.

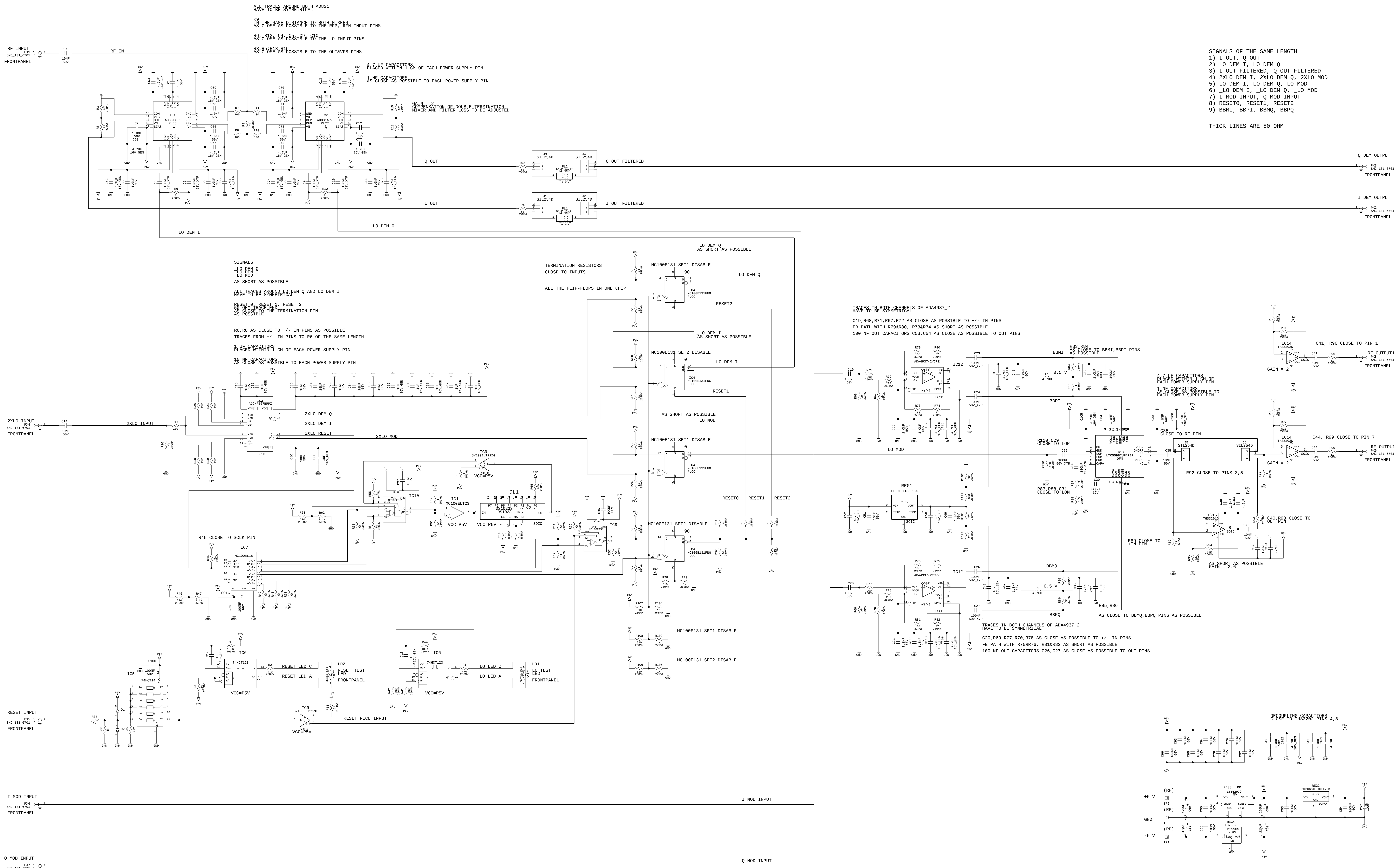
Appendix IV

Front panel



Appendix V

Schematic



- SIGNALS OF THE SAME LENGTH
- 1) I OUT, Q OUT
 - 2) LO DEM I, LO DEM Q
 - 3) I OUT FILTERED, Q OUT FILTERED
 - 4) 2XLO DEM I, 2XLO DEM Q, 2XLO MOD
 - 5) LO DEM I, LO DEM Q, LO MOD
 - 6) LO DEM I, LO DEM Q, LO MOD
 - 7) I MOD INPUT, Q MOD INPUT
 - 8) RESET0, RESET1, RESET2
 - 9) BBMI, BBPI, BBMQ, BBPQ

THICK LINES ARE 50 OHM

DECOUPLING CAPACITORS
CLOSE TO THS3202 PINS 4, 8

OŚWIADCZENIE

Oświadczam, że Pracę Dyplomową pod tytułem „A quadrature frequency converter in a feedback loop of high frequency cavities in the Proton Synchrotron at CERN”, którą kierował dr inż. Krzysztof Kurek, wykonałem samodzielnie, co poświadczam własnoręcznym podpisem

.....

Tomasz Truszczyński