ELECTRONICS FOR THE TTFL CAVITY-TYPE BEAM POSITION MONITOR

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Abstract

Single cylindrical cavities where built to detect the beam displacement inside each TESLA Test Facility Linac cryostat. The amplitude of the TM110-mode excited by an off-center beam yields a signal proportional to the beam displacement and to the bunch charge, its starting phase gives the sign of the displacement. This narrowband signal having a center frequency of 1.517 GHz is detected in a homodyne receiver. Two opposite cavity signals are combined in a stripline hybrid to reach the desired resolution of less than 10 microns. After filtering and amplification the output signal of the difference-port is mixed with the filtered sum-signal down to DC. Many of the RF-components were developed in-house. The paper describes the basic concept and most of the receiver components.

1 INTRODUCTION

Single cylindrical cavities will be used to measure the beam position at all superconducting quadrupoles in the TESLA Test Facility Linac (TTFL; see also [1] and [2]). Since there is one beam position monitor (BPM) in every cryostat, the first monitor is now in operation. Beside that, cavities working at room temperature having the same resonance frequency of 1.5167 GHz were built for the second bunch compressor (see also Fig.1).

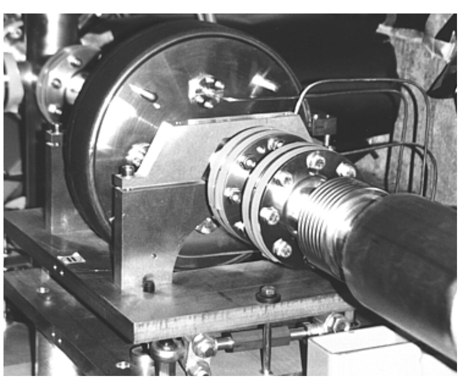

Figure 1: Warm monitor installed in the TTFL

The amplitude of the cavity TM_{110}-mode excited by an off-axis beam yields a signal proportional to the beam displacement and the bunch charge. Its phase relative to an external reference gives the sign of the displacement. Both TM_{110}-polarizations have to be measured to obtain the displacements in x and y, respectively.

The seventh harmonic of the bunching frequency of Injector I (216.7 MHz) has to be within the TM_{110}-bandwidth. This is very important for the cold monitors to avoid an active tuning system inside the cryostats. For the monitor installed in the first cryomodule, a frequency offset of $120\cdots140$ kHz has been measured after every cooldown. The temperature of the warm BPMs is stabilized in a thermostat and can be changed to tune the monitor slightly (about 20 kHz/K). In both cases CrNi was chosen as the cavity material to measure also individual bunches spaced at $1\ \mu$s (Injector II). The antennae are replaceable to allow a pre-tuning by adjusting the coupling. This paper describes the concept and all basic components of the detector electronics for the TTFL cavity BPMs.

Signals

For Injector I the cavity monitors monitors can measure only an average over the bunch train. The amplitude of the TM_{110} excited by a beam at a position δ_x can be estimated for a cavity without beam pipes [3]. The voltage at the output of the cavity into a $50\ \Omega$-system is than given by

$$V_{110}^{out}(\delta_x) = \langle I_b\rangle \cdot M_b \cdot \sqrt{\frac{R}{Q}\cdot Q_L \cdot \frac{\beta}{1+\beta}\cdot 50\,\Omega} \quad (1)$$

$$M_b = \frac{\delta_x \cdot a_{11}}{2J_1^{\max}R_{\mathrm{res}}} \quad \text{and} \quad \frac{R}{Q} \approx 130.73\cdot\frac{l}{R_{\mathrm{res}}}\cdot {T_{tr}}^2 \quad (2)$$

a_{11} is the first root of J_1, $\mathrm{T_{tr}}$ the transit time factor and $\beta = 1$ the coupling factor. The cavity radius R_{res} is about 115 mm, its length l 52 mm. Assuming a beam current $\langle I_b\rangle$ of 6 mA, critical coupling and a loaded Q_L of 1000, this yields a voltage of 1.5 mV for a beam offset of $1\ \mu$m. Even for a noise figure N_F of 10, this is clearly above the noise voltage V_n in a bandwidth of $B = 10$ MHz

$$V_n = N_F\sqrt{k_0\cdot B\cdot T\cdot 50\Omega} \approx 1.42\cdot 10^{-5}\ \mathrm{V} \quad (3)$$

k_0 is the Boltzmann constant, $T = 293$ K the temperature.

Since the field maximum of the common modes is on the cavity axis, they are excited much stronger than the TM_{110} by a beam near the axis. To reach the desired resolution of less than $10\ \mu$m, two opposite cavity signals are combined in a stripline hybrid. The rejection of common field components is limited by the finite isolation between the Σ- and the Δ-port of the hybrid. In addition, a frequency sensitive TM_{010}-rejection of more than 69 dB is required [2].

2 CONCEPT OF THE ELECTRONICS

The TM_{110}-amplitude is detected in a homodyne receiver by mixing the cavity output and a reference signal down to DC (Fig.2). For Injector I, this reference signal at the TM_{110}-frequency is phase-locked to the bunching frequency of the injector. When the beam is on the right, the system can be set up to give positive video polarity. The signal changes the phase by 180^o when the beam moves to the left, and for a centered beam it becomes zero. 12-bit ADCs are used to digitize this bipolar signal. Most of the components were developed in-house.

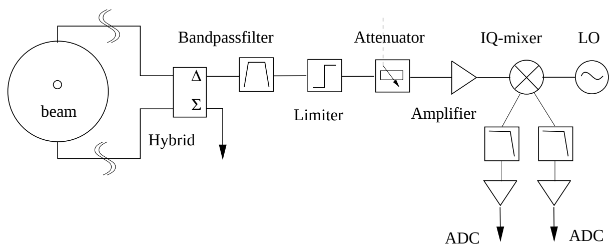

Figure 2: Signal processing scheme for TTFL cavity BPMs; only the y-channel is shown

The electronics reject all frequency components outside the $1-2$ GHz-band. But even in this band, the VSWR is larger than 1.0 and a part of the incoming signal will be reflected into the cavity. To reduce standing waves on the signal cables, attenuators were inserted in each input. Phase trimmers are inserted at the electronics front-end to adjust the physical length of all cables. The phase error remaining after calibration is less than 2^o.

2.1 RF-Module

The first component in the signal chain is a narrowband 180^0-*hybrid*, operating at the TM_{110}-mode frequency. Very important is the isolation between the Σ and Δ-port, which determines the resolution of the BPM. This component was realized by a commercial 90^0-hybrid and an additional delay line in one arm. The isolation of the whole hybrid is more than 32 dB, more than that of a standard (octave-band) 180^0-hybrids.

The commercial *bandpass filter* has a bandwidth of 120 MHz and a stopband rejection at 1.04 GHz (the frequency of the TM_{010}-mode) of more than -85 dB. Stripline-filters having similar passband characteristics were also developed. The main problem with this design was the stopband characteristic: Due to the design (side-coupled quarter-wavelength lines) there was a parasitic passband at 3 times the fundamental frequency, 4.5 GHz.

A coaxial limiter was inserted between the first passiv components and the rf-amplifier. It has to protect all following active devices against signals with higher input levels. The max. flat leakage of the selected device is sufficient to protect the rf-amplifier; the IQ-mixer is than protected by the rf-amplifier output level.

The measurement range - a fine range of better than 500 μm and a coarse range of up to 10 mm - is selected by the *digital attenuator*. Its value of 0 or 30 dB is remote-controlled by software. For the *rf-amplifier*, the

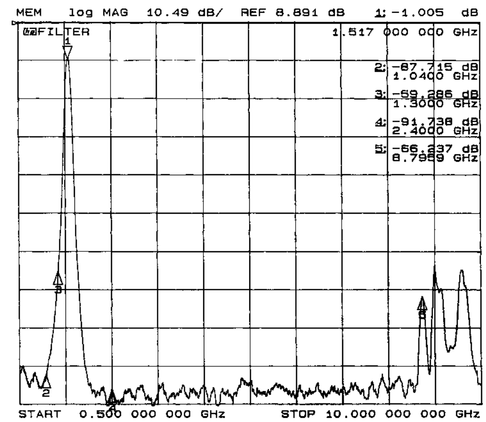

Figure 3: Bandpass filter with a center frequency of 1.517 GHz, measured insertion loss and stopband rejection

INA 10386 of Hewlett-Packard was used because of its gain ($+27$ dB) and a moderate noise figure. These amplifiers were installed in a balanced-amplifier scheme, as shown in Fig.4. The main reason for this realization is the improved VSWR, even at a reduced supply voltage. In addition, the upper limit of the linear amplification range is 3 dB higher than for a single amplifier.

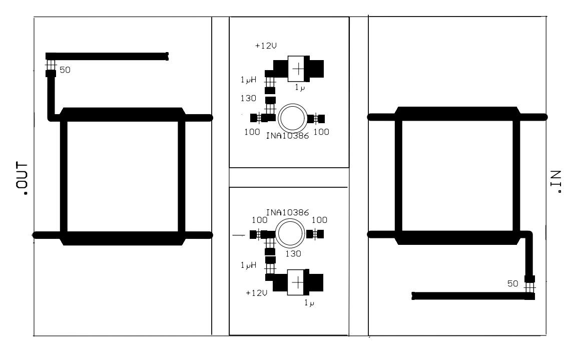

Figure 4: RF-amplifier - board layout and parts

For the position measurement the amplitude of the rf-signal at 1.5167 GHz has to be detected. Since this frequency is too high for a direct detection, the signal is down-converted to DC. Usually, this is done in a mixer using a local oscillator as a reference. The output signal depends on the phase difference between the two input signals.

Here, a so-called IQ-mixer (*In-phase/Quadrature*) scheme was realized, providing two equal amplitude IF outputs that are in phase quadrature. It consists of two mixers, a power splitter and a hybrid. Both input signals - *LO* and *RF* - are mixed in phase and in phase-quadrature. This results in a coordinate system, where the vector of the output signals I and Q is fixed for identical input frequencies. The design of these devices and an error correction/calibration are described in [4]. The advantage of an IQ-system is that for a constant signal level the measured voltage (vector) is fixed even if the phase difference varies.

The output signal of a single IQ-mixer channel has a maximum signal level of 0 dBm. To use the full ADC-range, this signal is amplified in IF-amplifiers, which reject also the higher harmonics.

2.2 LO-Module

The LO-system generates, amplifies and distributes the LO-signal at 1.5167 GHz along the machine. This reference signal is generated in an oscillator phase-locked to the 216.7 MHz-signal of Injector I, and amplified. A local LO-system in each of the electronics provides a constant reference signal (in amplitude) for each of the two IQ-mixers. The signal level required for their best perfomance is about $+10$ dBm for each channel, which includes already the 3dB-power divider inside.

Again, a balanced amplifier is used, but in this case it is saturated and the output signal is constant over an input range of more than 6 dB. The phase difference is not important, since this signal is used in a calibrated IQ-system.

3 CALIBRATION AND TESTS

Calculation of the Beam Position

Many calibration constants are needed to calculate the transverse position of the electron beam:

$$\delta x = U_{sa} \cdot \frac{1}{I_b} \cdot \frac{1}{k_{\text{cav}}} \cdot \frac{1}{k_{\text{cable}}} \cdot \frac{1}{k_{\text{el}}} \tag{4}$$

- U_{sa} is the calculated voltage of one channel (inlcuding the corrections for the IQ-mixers)
- I_b is the beam current measured in a current monitor
- k_{cav} is the excitation of the TM_{110}-mode in the cavity including the coupling of the antennae
- k_{cable} is the attenuation of the signal between the cavity and the electronics (including the attenuator)
- k_{el} is the amplitude calibration of the electronics.

Following eqn.(1), the constant k_{cav} can be calculated from

$$k_{\text{cav}} = \sqrt{\frac{\beta}{1+\beta}\frac{R}{Q} \cdot Q_L \cdot Z_0 \cdot T_{\text{tr}}\frac{3.293}{R_{\text{res}}}} \tag{5}$$

Calibration

The electronics for two cavity monitors have been installed, all cavities and cables (e.g. attenuation, physical length) were tested and calibrated. Beside that, a method described in [4] was applied for the calibration of each signal channel. It is based on the idea, that the IQ-mixer should measure a rotating vector for a cw-input signal of constant amplitude, which phase with respect to the LO-signal varies. All calibration parameters for each of the IQ-mixers were caculated by an optimization program. The remaining error is less than 5% for a phase variation of 360^o. This calibration includes also the two ADC-channels which are required for each of the cavity planes.

A typical result of a linearity test is shown in Fig. 5. The phase errors introduced by the step attenuators (used to change the input signal level) can be neglected after the calibration. The 1 dB-compression point is about 700 mV, the dynamic range more than 35 dB.

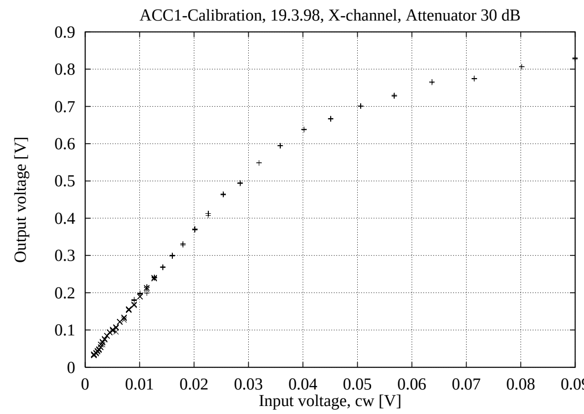

Figure 5: Linearity of the electronics measured after the calibration. The 30 dB-attenuator was switched ON.

For the automatization of this procedure, a calibration module has been built recently. It mainly contains an amplifier, an attenuator and a remote-controlled phase shifter. For the latter, an IQ-mixer developed for the electronics was calibrated and used as a vector modulator. In the latest modification of the electronics, the RF-module and the LO-module are on a single board.

4 ACKNOWLEDGEMENTS

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5 REFERENCES

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