

## CONTROL SYSTEM OF RF STATIONS FOR NICA BOOSTER

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### Abstract

NICA (Nuclotron based Ion Collider fAcility) is an accelerator complex, which is being built in JINR (Dubna, Russia). The system described in this paper is controlling the RF stations of Booster, the first element of the NICA complex. Control System consists of two devices: Intellectual Controller and Tester module. The first one is designed for precise measurement of magnetic field, generation of the acceleration frequency in accordance with measured field and control of RF power and pre-amplifiers. Intellectual Controller is a real-time feed-forward system with 40  $\mu$ s loop time. It is based on ARM microcontroller and bare-metal control programs are used to reach maximum performance. Approaches that were used to achieve maximum performance are elaborated and presented in this paper. The second part of system - Tester is a simulator for tuning and checking the RF stations before start of operations or in absence of real accelerator. The achieved accuracy in chain 'magnetic field' - 'acceleration frequency' is better than  $5 \cdot 10^{-5}$ . Plans on feedback incorporation to stabilize ion beam behavior via frequency and phase tuning are discussed.

are measured and set, they are presented at table 1. Both RF Controller and Tester modules use telnet over Ethernet interface for interaction.

Table 1: Input and Output Signals of Controller

Signal	Channels	Rate	Resolution
Output signals			
Master frequency	2	25 kHz	24
V cavity	2	25 kHz	12
I anode	2	100 Hz	10
Input signals			
Field sensor	1	25 kHz	18
V cavity	2	25 kHz	12
Phase difference	1	25 kHz	12
Phase bias	1	25 kHz	12
Frequency bias	1	25 kHz	12
V preamplifier	2	100 Hz	12
I anode	4	1 kHz	12
V rectifier	6	1 Hz	12
V filament	2	1Hz	12
Synchronization	7	N/A	N/A

### INTRODUCTION

The superconducting accelerator complex NICA (Nuclotron based Ion Collider fAcility) is constructed in JINR, Dubna. Project aims to provide collisions of heavy ion beams in the energy range from 1 to 4.5 GeV/u at the luminosity level of  $1 \cdot 10^{27} \text{ cm}^{-2} \cdot \text{s}^{-1}$ . The Booster is a cycling accelerator of 197Au<sup>32+</sup> ions and serves as one of the elements in the injection chain. It accepts particles with energy 6.2 MeV/u, accelerates them and extracts with energy 600MeV/u to the next acceleration stage - Nuclotron [1].

RF subsystem of the Booster consists of: two cavities, two RF stations, intellectual controller and tester module. RF stations include power amplification cascades and low-voltage electronics. They provide 10 kV acceleration voltage in required frequency range (0.5-5 MHz) on cavity gaps [2]. Primary function of the controller is to generate master frequency depending on the current value of magnetic field. Inaccuracy of frequency setting must not be worse than  $\pm 2 \cdot 10^{-5}$ .

### RF CONTROL STRUCTURE

RF control structure is shown on Fig. 1. Signals from induction sensor and synchronization pulses are passed through the Tester module to Controller. Tester module allows to substitute real signals with imitated scenario or to through-pass these signals. Frequency, phase and cavity voltages are adjusted according to measured magnetic field and acceleration stage. Other technological parameters

### Controller Structure

Observing table 1 one may conclude that controller module must supply 4 fast (25 kHz) output channels and measure 6 signals with the same rate. Most calculations and data conversions will require floating-point arithmetic. It is also necessary to arrange data interchange between several independent measurement and control devices. E.g. to generate master frequency controller must receive data from measuring ADC, convert, correct and integrate the resulting value to attain field value then calculate frequency, apply necessary corrections, then convert it again and finally send this data to DDS. At the same time other signals must be changed and measured: cavity voltages, cathode currents, preamplifier voltages etc.

Probably betatron and synchrotron oscillations damping would be necessary for Booster. It is very hard to predict feedback parameters before Booster starts operation. Therefore we decided to implement feedback later using separate module which will provide frequency and phase bias analog signals.

Nowadays there are two main approaches for creation of such complex high rate systems: FPGA and high-end micro-controllers. Our reasons for choosing micro-controller were following: the need of floating point arithmetic, complex logic with different modes of operation that probably will require expansion and finally the need to provide TCP/IP interface for the interaction with Booster control system. At

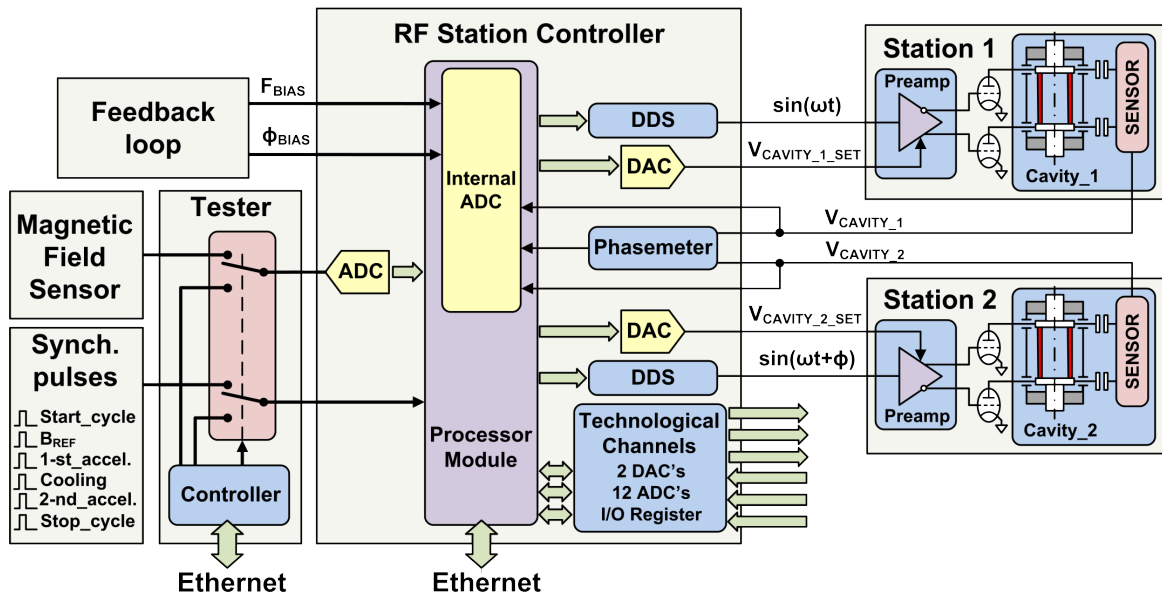


Figure 1: Radio frequency control structure.

the beginning of our work we planned to have Linux with tango-server on-board. Unfortunately we couldn't achieve necessary performance with any variation of Linux, so bare-metal approach was used.

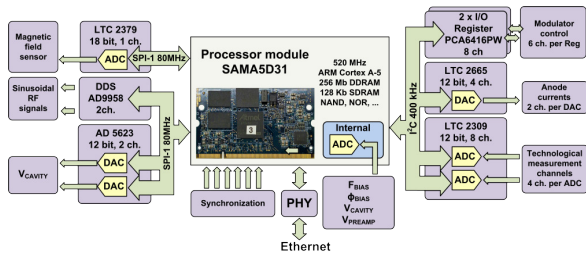


Figure 2: Internal structure of the controller.

Structure of the controller is shown on Fig. 2. We used Atmel SAMA5D31 micro-controller. Availability of on-board 1 MHz 12 bit 8-channel ADC was a significant advantage. Usage of SAMA5D31-CK SO-DIMM board allowed us to simplify PCB design and accelerate production. Availability of different interfaces (I2C, SPI, UART, Ethernet) on this controller was a huge plus. It must be noted though that we had to give up usage of any (even real-time) operational system to provide necessary reaction rate with  $40\mu s$  cycle. Apart from such bare-metal approach we had to thoroughly optimize our program code. Resulting algorithm is shown on Fig. 3.

Main cycle is realized using three routines. One of them is called "Fast cycle". It is initiated every  $40\mu s$  by high-priority timer IRQ. Signal from inductive sensor is measured, derivative of frequency is calculated and passed to DDS. Fast signals (look at Tab. 1) are measured by ADC and generated by DAC's. Another one is "Synchronization" state machine. This state machine is operated by synchronisation I/O that generates low-priority IRQ. Last one is a "Slow cycle" which

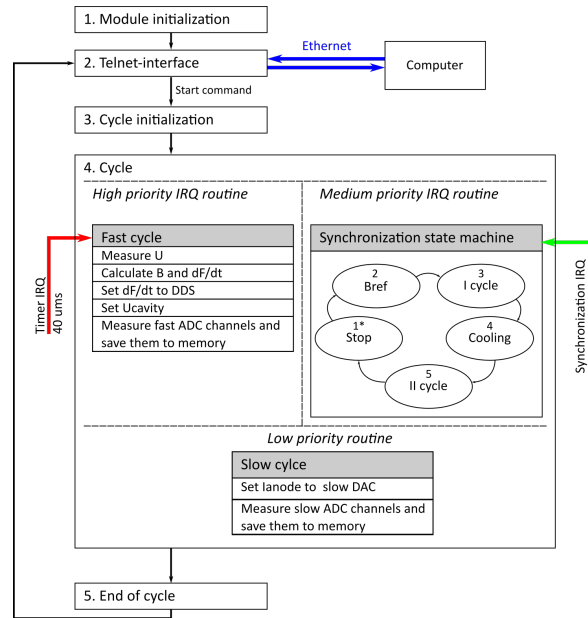


Figure 3: Program structure.

is a loop that measures and generates slow signals. This code works between the other two and thus strict synchronization couldn't be achieved, fortunately it isn't necessary.

Another restriction we had to establish is following: interaction with Booster control system using Ethernet protocol is available only between cycles, while in-cycle Ethernet controller is turned off. Apparently an ideal approach in such case would be the usage of 2-core microcontroller, or SOC with FPGA and ARM kernel e. g. CYCLONE V SOC. But when we started this work, boards like MITYSOM-5CSX on SODIMM were still unavailable. And PCB layout of CYCLONE V SOC is quite problematic in itself and was considered more tedious than program optimization.

Both tester and controller modules are managed over the Ethernet interface using text-based command protocol over telnet. RS-232 interface is provided for reprogramming and debug. Apart from adequate embedded programs, testing software was developed. This software will also serve as a reference for implementation of corresponding modules in NICA Booster control system which is based on TANGO [3].

### Low Level Frequency Generation

Main task of the controller is to provide low-level sinusoidal signal with frequency in accordance with the magnetic field. The dependency can be presented in the following form:

$$f(B) = K \frac{aB}{\sqrt{b + CB^2}}, \quad (1)$$

where  $K$  - is the harmonic number,  $B$  - is magnetic field, and  $a, b$  and  $C$  are coefficients. We'll use Eq. 1 from now on.

Most common approach for the low-level RF control systems nowadays is to generate sinusoidal signal using DDS (direct digital synthesis). Both specialized IC's and FPGA with DAC implementations are used. We decided to use latter because our system is based on microcontroller. To alleviate possible difficulties with inter-channel synchronisation two-channel AD9958 chip was chosen.

Actual signal from the inductive sensor is proportional to the change of magnetic flux over time. If the sensor is stationary, then the area is constant and signal  $U$  is proportional to the derivative of the magnetic field:

$$U = \frac{1}{\alpha} \frac{dB}{dt}, \quad (2)$$

where  $\frac{1}{\alpha}$  is a proportionality constant.

Therefore to calculate  $B$  and for Eq. 1, one has to measure the following integral:

$$B(t) = \alpha \int_0^t U dt. \quad (3)$$

We can use several methods to generate master-frequency that varies according to the law Eq. 1. First of them is not to measure integral at all and work by a table instead. Actually if you have perfectly repeatable magnetic field source and perfect synchronisation, you can calculate values of  $f(t)$  beforehand, put them in a table and generate them using DDS at appropriate moments of time. We can note that this approach doesn't require measurement of magnetic field at all. The downside of this approach is the complexity of creating sufficiently stable magnet power supply and providing necessary synchronisation accuracy [4].

Rather wide-spread method to measure integral at Eq. 3 is based on the use of voltage to frequency converters (VFC) followed by counter [5].

Other way to determine integral is to measure inductive sensor output voltage (i.e. derivative  $dB/dt$ ) with the help of fast, precision ADC and then to sum the ADC samples in order to calculate the integral value. This method may be called "direct digital integration" [6]. Usage of the fast

precision ADC allowed us to use a specific method of frequency tuning to achieve higher accuracy. This method is described below.

### Our Method of Frequency Tuning

As described previously, one of the approaches is to acquire field value, substitute it to the Eq. 1, integrate it, apply necessary corrections (e.g. feedback signals, etc) and set calculated frequency value to DDS generator. This was the approach that we realized in our first controller prototype.

The main consideration on this approach (besides accuracy of measurement for which appropriate means were taken) is its "stepping" nature, which introduces a quantization error. We can estimate this error in the following way. Highest frequency change rate is 2.25 MHz/sec. With practically achievable cycle time  $\Delta T = 40 \mu s$  we get 25 Hz quantization error. That is around  $5 \cdot 10^{-5}$  at the starting frequency 500 kHz.

Now we would like to describe the method which allows to simplify calculations, lower quantization error and loosen cycle time requirements. This method relies on the ability of DDS to perform linear sweep of frequency with variable sweep rate over time. This ability is implemented as shown at Fig. 4. Internal down  $\delta t$  counter operates at DDS clock (25 MHz in our case), when it reaches zero a value written in  $\delta f$  register is added to current frequency register and counter is reset to initial value. Current frequency register is used by sinusoidal waveform generator to provide output signal. Essentially this ability means that we are able to provide DDS with frequency derivative  $\frac{df}{dt}$  modifying both difference at each step  $\delta f$  and stepping rate  $\delta t$ . If  $\delta t$  is lower than the measurement/integration cycle time  $\Delta T$ , we get an advantage in precision.

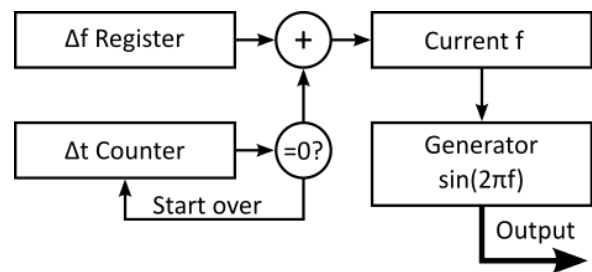


Figure 4: Simplified DDS frequency sweep implementation.

The illustration to this fact is provided on Fig. 5.  $U(t)$  is a typical signal from induction sensor,  $B(t)$  is an appropriate integral. Let's assume that our cycle period is  $\Delta T$ , therefore individual readings are taken at moments of time  $t_0, t_1, \dots, t_i$ , with  $\Delta T = t_i - t_{i-1}$ . Graph  $f(t)$  represents the approach when on each step we measure  $U(t)$ , calculate  $f(\alpha \int U(t_i))$  and provide DDS with new frequency value. Graph  $f_{sweep}(t)$  represents an approach with derivatives which will be described in detail below. Apparently this approach allows to significantly lower quantization error.

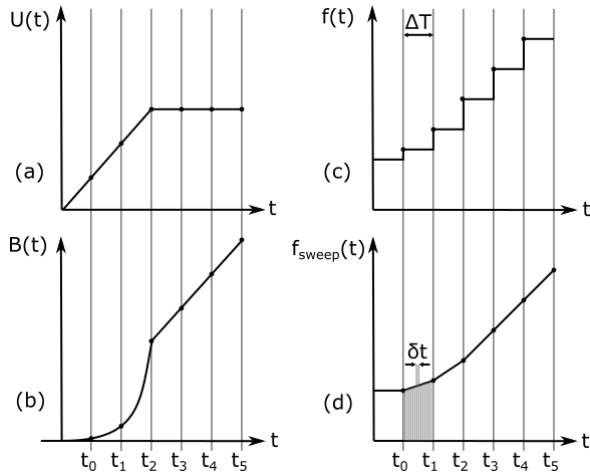


Figure 5: Demonstration of sweep approach superiority. (a) – inductive sensor signal  $U(t)$ ; (b) – magnetic field  $B(t)$ ; (c) – frequency form for “stepping” approach  $f(t)$ ; (d) – frequency form for “sweep” approach.

To calculate  $\frac{df}{dB}$  we start from the Eq. 1 and take its derivative over  $dB$ :

$$\frac{df}{dB} = \frac{Ka}{\sqrt{b + CB^2}} - \frac{KaCB^2}{(b + CB^2)^{3/2}}, \quad (4)$$

after examination of Eq. 4 taking Equations 1 and 3 into account we get the following:

$$\frac{df}{dt} = \frac{df}{dB} \frac{dB}{dt} = \frac{f}{B} \left(1 - \frac{C}{K^2 a^2} f^2\right) \alpha U. \quad (5)$$

For the actual calculations following discrete form is used:

$$B_i = B_{i-1} + \alpha U \Delta T \quad (6)$$

$$\delta f_i = \frac{f_{i-1}}{B_{i-1}} \left(1 - \frac{C}{a^2 K^2} f_{i-1}^2\right) \quad (7)$$

$$f_i = f_{i-1} + \delta f_i \delta t \quad (8)$$

Notable feature of these equations is the absence of square root operation. Values  $f_0$  and  $B_0$  could be calculated before the start of an acceleration cycle.

Currently we are using the described method with  $\delta t = 4 \mu s$ , therefore achieving theoretical frequency error level of  $5 \cdot 10^{-6}$  in the worst case. Error measured using a frequency meter is  $1.4 \cdot 10^{-5}$  and is determined by different measurement and synchronisation inaccuracies.

## TESTER

Tester Module consists of digital control unit and analog Front-End. Analog Front-End is based on DAC8831 IC with 16 bit resolution. The core of digital control unit is LPC2478 processor, which provides DAC control of analog Front-End, synchronization pulses generation, multiplexer control and Ethernet interface for inter-operation with Booster control system. Particular attention was paid to analog Front-End circuit design. The block diagram of analog

Front-End is shown in Fig. 6. Front-End is a combination of two OPA2277 operational amplifiers. First one provides gain and offset for DAC output signal. Second generates  $\pm 1$  V signal on 120 Ohm input resistance of the Controller module. Using offset adjustment circuit, high precision resistors 0.1% with  $5 ppm/^\circ C$  temperature coefficient and precision voltage references (LTC6655) allow to achieve high accuracy of the magnetic field derivative. Resulting noise integral of magnetic field simulation is less than  $10^{-5}$ .

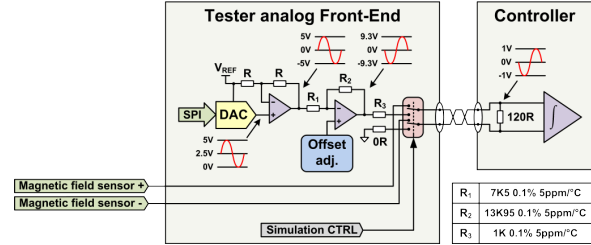


Figure 6: Tester structure.

Ethernet interface for inter-operation with booster control system.

## CONCLUSION

Controller and Tester modules were designed manufactured and tested in working conditions with RF stations generating high voltage. Usage of SAMA5D31-CK board allowed to significantly reduce prototyping time. New approach for frequency tuning was devised. Overall measured inaccuracy using frequency meter is better than  $1.4 \cdot 10^{-5}$ . After NICA Booster commissioning beam feedback module would be produced and incorporated to the LLRF control system. All devices provide simple telnet interface for integration in NICA control system.

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