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ELECTRONICS FOR THE CONTROL OF THE RF SYSTEM*

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This note describes the operation of the major components used for controlling the phase and the field level of the PEP RF cavities. The block diagram of one RF station is decomposed into several control loops: Each cavity has a tuners' servo loop which maintains the frequency constant and also keeps the field of each cell at the same level; the total gap voltage developed by a pair of cavities is obeying the command of the gap voltage controller; finally, the phase variation along the amplification chain and the klystron are compensated by a phase lock loop.

The design criteria of each loop are set forth and the circuit implementation and test results are presented.

The purpose of this report is to acquaint interested people with the design philosophy and to allow them to evaluate the capabilities of this system and its behavior during operation of the machine.

1. Cavity Tuners' Servo

Two types of perturbations are dealt with here: one is the detuning of the cavity due to thermal expansion and to reactive beam loading, the other is an instability of the cavity brought about by an uneven

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temperature distribution which can result in a run away condition, where more and more of the stored energy concentrates in one end of a cavity.

The first perturbation is stabilized by a feedback loop controlling the common penetration of the tuners. The second perturbation can be minimized by the differential positioning of the tuners, either in cells 2 and 4 or in cells 1 and 5. Figure 1 is a block diagram of the cavity tuners' servo. A phase detector is sensing the phase difference between the input field and the field of the center cell; after suitable amplification this signal is applied in a common mode fashion to both tuner drivers.

Two peak detectors measure the field levels of the outboard cells and their difference is also applied to the tuner drivers, but this time the signal is connected in a push pull fashion.

1.1 Design of the phase control loop. It is based on the data of Fig. 2 which represents the cavity detuning as a function of the tuner's travel. Between start-up, with a cooling water temperature of 70°F, and full power, with a cooling water temperature of 95°F, we have measured a total detuning of 300 kHz and a tuner's travel of 70 mm.

Let us assume that the phase across the cavity is to be held constant within 0.2° over this range of operation; the phase detector signal which corresponds to 0.2° is 5.4 mV and it must be amplified so as to produce a tuner motion of 70 mm. The phase detector will be described subsequently in this report. Since 7V are required at a tuner driver input to get such a displacement, the amplifier gain is found to be $(7:5.4) 10^3 = 1300$.

Fig. 1 Block diagram of the cavity servo loops.

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Fig. 2 Cavity tuning curve for two ganged tuners.

The resolution of the tuner's positioning servos is determined in the following way: the cavity 3 dB bandwidth being taken as 17 kHz, with the help of the curve of Fig. 2, one finds that a phase variation of 90° is obtained with a tuner travel of 2 mm. We desire a resolution of $\pm 0.2^\circ$; this corresponds to a tuner motion of ± 0.004 mm. This resolution has been achieved by picking a lead screw having a pitch of 1.25 mm per turn and a stepping motor requiring 400 pulses per revolution (half step mode).

1.2 Design of the field balance loop. The data used for this design are presented on Fig. 3; it is a plot of the cells' field level as a function of the differential position of the tuners, the frequency being kept constant. We note that a 5 mm differential position of the tuners yields a ratio of 0.5 for the field levels of cells 1 and 5, i.e., 0.5V at the drivers input yields 5V at the field balance detector for the maximum power level. The field detectors will be described in the following chapters; they are calibrated to give a maximum signal of 10V for 50 kW per cell. The gain from the tuner drivers to the field balance detector output is therefore 10. For a control of the fields within 1%, the loop gain must be increased by another factor of 10. In practice, the smallest field unbalance we have achieved was determined by the cable and field detector's unbalance.

1.3 System dynamic response. A block diagram of the tuners' positioning servo is shown on Fig. 4. The voltage-to-frequency converters are set for a maximum frequency of 800 Hz. This yields a top speed of 120 rpm in response to large inputs. However, for small signals we

Fig. 3 Field profile of the five cavity cells as a function of the tuners' insertion.

- a) Tuner #1 inserted 10 mm, tuner #5 withdrawn 10 mm.
- b) Tuner #1 inserted 5 mm, tuner #5 withdrawn 5 mm.
- c) Flat field.

Fig. 4 Block diagram of a tuner positioning servo.

have measured a bandwidth of 3 Hz, and under the control of a manual potentiometer the tuner reacts pretty much within the speed of human response.

It was found necessary to slow down the response of the field balance loop; indeed, during the initial cavity tuning process, large phase and amplitude differences are detected which interfere with the transients damping. Various interlocks insure a safe operation of the tuners; the most important is perhaps the parking feature: in the event of an RF trip-off, the tuners are given a command to move toward a preset position; this position is programmed by the operator and must be picked so that the cavity be not excited by a circulating beam.

2. Phase Stabilization

The PEP klystrons are operated with beam voltages between 40 and 63 kV and exhibit typical phase shifts of 390° over this range. In addition, small phase variations can be expected from other components along the amplification chain and along the waveguide as a result of RF level adjustments or temperature changes. The phase feedback loop has been designed to maintain a phase stability of the cavity fields within $\pm 1^\circ$ over several hours of operation, and at any power level. Figure 5 is a diagram of the phase stabilization circuits. A sample of each cavity field is taken from the center cell and transmitted via 100 feet of phase stable cable* to a summing hybrid junction, in order to produce the vector sum of two nearly in-phase signals. Consequently, the phase

* Andrew type LDF 4-50, with a phase temperature coefficient of approximately $7 \text{ ppm}/^\circ\text{F}$.

Fig. 5 Circuit of the phase control loop.

stabilization is being made on the sum of the cavities' field vector and this arrangement allows small individual trimmings of the cavities phase offsets (see Fig. 5) without altering the phase of the total gap voltage. The amplified phase error is added to some variable offset voltage and applied to the control port of the phase shifter. This offset is required to insure that upon closure of the loop, or upon turn on of the RF, the phase shifter operating point is compatible with the klystron selected anode voltage. Indeed, this system must accommodate a large range of phase shifts, and the initial locking can occur around two stable points separated by 360° . To remove this ambiguity the offset voltage is made to track approximately the klystron phase curve, and is simultaneously programmed with the klystron anode voltage.

2.1 Phase detectors. For this type of system the phase detection must introduce errors smaller than $\pm 1^\circ$ over a power range of 30 dB. Several approaches exist to make such measurements. One circuit which we call method A^[1] consists in the linear detection of the vector sum of the two RF signals. To make the output immune to low frequency drift, the RF reference is made to "wobble" in phase $\pm 90^\circ$ at the rate of a low frequency square wave thereby making the phase detection an AC measurement; the phase information is contained in the peak to peak voltage of a similar square waveform. Method B^[2] converts both RF signals to some intermediate frequency by mixing. The IFs are then limited in amplitude and the phase measurement becomes simply a duty factor detection on the output of an EXCLUSIVE OR gate which responds to the two IF signals. Method C^[3] uses also a down-frequency conversion of the RF signals but this time by means of a sampling technique;

the rest of the detection is analogous to the one applied in method B. Any of these three methods could possibly be adapted to our requirements; method A was discarded because of its narrower phase range and its sine function type of characteristic. The superiority of method C over method B, yet a minor one, is that the frequency conversion by sampling can be made 100% efficient and permits an accurate amplitude measurement as a by-product. Yet, for these two types of detectors the difficulty is not found in the down-frequency conversion but rather in the design of a limiter free of phase shift, and an output circuitry with no DC drift. Indeed, a measurement within $\pm 1^\circ$ is equivalent to errors smaller than $\pm 0.3\%$, since 1° represents 1 part in 360, and this goal calls for some care in components and circuit selection.

Figure 6 is a layout of the phase detector scheme. Two identical sampling heads convert 353.210 MHz into an intermediate frequency of 100 kHz when the sampling frequency is adjusted to be 5.044 MHz. This is equivalent to beating the 70th harmonic of the sampling frequency with the RF, except that the efficiency can be made 100% by means of positive feedback; the two IF signals have the same phase difference as their respective RF inputs and also the same amplitude. This conversion process is linear for power level from + 10 dBm to - 60 dBm. The AGC limiters use an MC 1590 variable gain amplifier which maintains constant outputs of ± 50 mV over a range of 60 dB. Beyond this point the circuitry is straightforward. An RS flip-flop connects alternately the output of a 30 mA current source to the outputs of two 15 mA sources; the rate is 100 kHz, however the duty factor is linearly related to the phase of the two RF inputs. The phase signal can be used as a current

Fig. 6 Principle of the phase detectors.

source of ± 15 mA for $\pm 180^\circ$, or transformed into an equivalent voltage source of ± 5 V. The dotted part of Fig. 6 represents the generator of the sampling frequency; this circuit phase-locks the IF signals to a reference oscillator. We use one such circuit for the synchronization of 12 phase detectors.

Test results: A production of 40 phase detectors was evaluated. Our criterion of $\pm 1^\circ$ error could be met by all units for power levels between + 10 and - 30 dBm. Half of this production could operate over a power range of 50 dB, and few units over 60 dB; in all cases the deterioration of the specifications could be imputed to the limiters which distort slightly the waveforms for high input levels, or simply exhibit a reduced operation range. A temperature variation of 30°F did not cause excessive phase errors.

2.2 Phase shifters. Electronic phase shifters are used in three functionally different places in each RF station:

- a. The klystron phase shifter is the control element in the klystron phase loop. The main requirement for this phase shifter is a linear response of phase versus control voltage. A large range of 540° is required since this phase shifter compensates for the large phase change the klystron exhibits when its high voltage is varied between 40 kV and 63 kV.
- b. Nulling phase shifters are used in conjunction with each phase detector to set the operating point of the klystron phase loop and the two tuner control loops. These phase shifters have a range of 180° each and are required to hold

Fig. 7 180° phase shifter module

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Fig. 8 Phase-shifter response and insertion loss as a function of bias voltage.

their phase shift, once set, to a constant value independent of ambient temperature changes.

- c. The station phase shifter is inserted between the RF reference line^[4] and an RF station and sets its phase in reference to the master oscillator. It has to have a range of 360° with linear response to its control voltage and good temperature stability,

All these requirements are met in the basic design of a 180° phase shifter module which is then cascaded as required for more range. The 180° phase shifter module is laid out in a microstrip line circuit with two voltage variable capacitance (VVC) diodes as control elements (Fig. 7). A detailed description of the module can be found in PEP-NOTE-283.^[5]

The typical insertion loss for a 180° phase module is 0.4 dB and the temperature coefficient 0.04 degrees of phase/ $^{\circ}\text{C}$. The typical phase shift as a function of the control voltage for the 360° station phase shifter is shown in Fig. 8.

2.3 Stability of the phase control loop. The choice of the frequency bandwidth for this loop depends on the range of synchrotron frequencies of the different machine configurations. It was deemed safer to make the klystron phase loop somewhat insensitive to the synchrotron frequency and since the lowest expected value for this frequency is 1 kHz, the open loop unity gain has been set below 1 kHz. The filters and loop amplifiers circuits are shown on Fig. 9 and their frequency response is plotted on Fig. 10. Since the combined transfer function of the phase

Fig. 9 Filters and amplifiers for the phase control loop.

Fig. 10 Open loop frequency response of the phase control.

detector and the phase shifter is approximately unity, Fig. 10 represents nearly the open loop gain characteristic. The various networks of Fig. 9 are such as to produce a minimum phase shift function and a 0 dB cross-over frequency, independent of the gain adjustment. The limiting feature at the output of the switched gain amplifier insures that upon closure of the loop, the phase shifter transient excursions remain within the acceptance window of the phase detector, that is $\pm 180^\circ$.

3. Amplitude Stabilization

A block diagram of the gap voltage feedback loop is shown in Fig. 11. The field levels in each cell of two cavities are measured by the RF amplitude detectors and then added up to get a total gap voltage. This signal is compared to an adjustable DC reference signal. The error, after amplification, is applied to a variable attenuator, which controls the RF drive power to the klystron.

3.1 RF amplitude detectors. Hot carrier diodes are used as peak detectors (Fig. 12). The diode charges a capacitor to the peak RF voltage minus a small voltage drop across the diode. The same type of diode is used in the feedback path of a unity gain buffer amplifier and thus compensates for the voltage drop across the detector diode. This feedback diode also cancels the strong temperature dependence of the diode voltage drop, since it is mounted in the same thermal environment as the detector diode. The resultant output of the circuit tracks the peak RF voltage over a range of 40 dB from a maximum of 10V down to 100 mV, where the error increases to about 10%. The field sampling

Fig. 11 Block diagram of the amplitude feedback loop.

Fig. 12 RF amplitude detector circuit diagram.

probes are set during installation to couple 1 watt for 50 kW dissipated in each cavity cell, and the detector is calibrated to give 10V output for a 1 watt input power.

An important part of the accuracy of the field measurement by this detector is the error introduced by standing waves on the line between the detector and the sampling loop. Because of the low source impedance of the sampling loop any reflection at the detector will affect the signal flowing from the sampling loop to a varying degree, dependent on the length of line between the loop and the detector. This source of error can be minimized by positioning the detector diode a number of half wavelengths away from the sampling loop. At any half wavelength point the RF voltage is largely independent of standing waves and directly related to the source voltage. For this reason, all components connecting the detector to the sampling loop were made with a specified electrical length, including the approximate 100 ft. length of coaxial cable.* A 400 MHz low pass filter is inserted in the line to prevent passage of beam induced higher order mode frequencies which could influence the detected field level. The sampling loops are installed in the cavities with an accuracy from loop to loop of $\pm 1.5\%$. Taking the other sources of error into account the accuracy with which the field level can be measured is $\pm 5\%$.

The cavity gap voltage is derived by adding the 5 cell voltages of a cavity. This gap voltage is set to match the calculated gap voltage

* Andrew type LDF 4-50.

for a given power dissipated in the cavity. This calibration will later be checked against the value deduced from the measurement of the beam lifetime.

3.2 Variable attenuator. The variable attenuator is used as the control element in the gap voltage feedback loop, as well as the element setting the minimum power level at which the phase control loop can tune the cavity to resonance. The drive power to the klystron in the tune mode can be as low as -30 dB from the saturating drive power. A variable attenuator was therefore designed with a 40 dB range and small phaseshift over its attenuation range. The microstrip design is shown in Fig. 13. Two PIN diodes^{**} vary their individual resistance to RF from $> 10 \text{ k}\Omega$ to 1Ω for bias currents of zero to 50 mA. For zero bias all the input power is divided by the 3 dB hybrid^{***} and transmitted past the PIN diodes to be absorbed in the 50Ω terminations. This is the maximum attenuation state. The signal emitted out of the isolated port of the hybrid in this state, can be adjusted to -40 dB by optimizing the match of the terminations, using little tabs on the microstrip line. With some bias current flowing through the PIN diodes, some power is reflected and adds up in the output port of the attenuator. At a bias current of 50 mA each, the PIN diodes act like short circuits across the lines and all the power is reflected and combined at the output port. This is the minimum attenuation state. The phaseshift

^{**} HP 5082-3040.

^{***} Anaren type 180260-3.

Fig. 13 Variable attenuator circuit diagram.

through the attenuator over most of its range is negligible, except in its high attenuation state where reflections from other points than the PIN diodes can add up with any phase. This condition can also be brought under control by properly attaching tabs to the microstrip line when optimizing the attenuation. By these means the phase shift of the attenuator could be held to within $\pm 10^\circ$. The attenuation as a function of bias current is shown in Fig. 14. It is very nonlinear. Since the attenuator is part of a feedback loop, which regulates the RF voltage in the cavity, a linear response between the RF output voltage of the attenuator and a control voltage of 0 to 10V was desired. A driver circuit was designed using a multifunction converter[†] and the nonlinear impedance of diodes, to linearize the overall response. Figure 15 shows the driver circuit and Fig. 16 shows the RF output voltage as a function of the control voltage of the attenuator-driver combination. The insertion loss at minimum attenuation is 0.8 dB.

3.3 Amplitude loop operation. The klystron, which is part of the amplitude feedback loop, works with highest efficiency when saturated with drive power. The operating point for a certain gap voltage can be set by adjusting the klystron high voltage to a level where the RF drive power is close to the saturation level, or the variable attenuator close to its maximum output. A feedback gain of 10,000 is then necessary to keep the error below 1%.

The unity gain bandwidth was set to 1 kHz to assure that the loop be insensitive to synchrotron oscillations.

[†] Burr Brown #4302.

Fig. 14 Attenuator response as a function of bias current.

Fig. 15 Circuit of the attenuator driver.

Fig. 16 Overall response of the attenuator with its driver, as a function of the control voltage.

4. Conclusion

The control of the PEP RF system has been implemented in a modular format allowing conveniently for modifications or maintenance. All analog input functions are applied either via a computer controlled Camac system, or directly at the RF station from a manual control panel. The monitoring signals are also made available to the computer system or can be observed on a variety of meters and monitoring jacks. Therefore each RF station can be entirely operated on the spot when there is no stored beam. However, phasing of several stations, or phasing of more than one region (a region comprises four or more stations), can only be done from the main control room via the PEP control system.

The circuits presented in this report have all been proved in their manual mode on the cavity test bench. To date, four stations have been run simultaneously in the field; the operation was very smooth and confirmed the soundness of the design.

This project has required an interaction with all members of the PEP RF group. We want to thank Matt Allen for his leadership of this group.

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