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# CONTROL ELECTRONICS OF THE PEP RF SYSTEM*

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## Abstract

The operation of the major components used for controlling the phase and field level of the PEP RF cavities is described. The control electronics of one RF station is composed of several control loops: each cavity has a tuners' servo loop which maintains the frequency constant and also keeps the fields of each cavity balanced; the total gap voltage developed by a pair of cavities is regulated by a gap voltage controller; finally, the phase variation along the amplification chain, the klystron and the cavities 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.

## 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 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 (Fig. 1). The design of the tuning control loop 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.

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 ±0.2°; this corresponds to a tuner motion

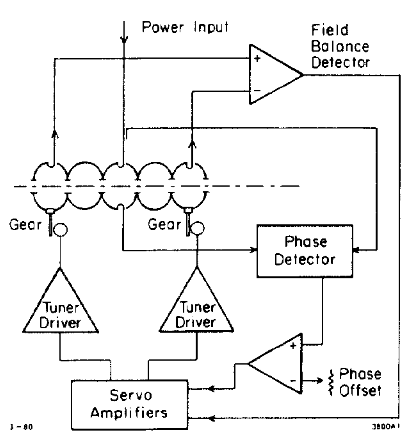


Fig. 1. Block diagram of the cavity servo loops.

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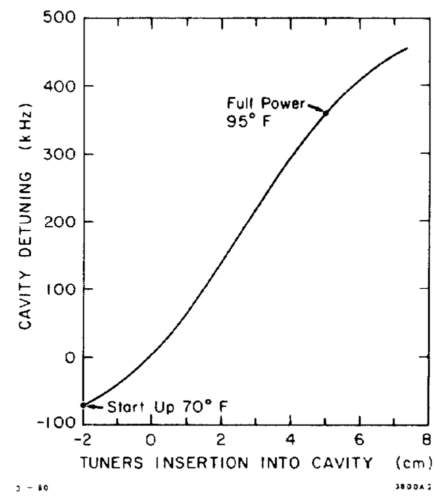


Fig. 2. Cavity tuning curve for two ganged tuners.

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).

The design of the field balance loop uses the data presented on Fig. 3; it is a plot of the cell's field level as a function of the differential position of the tuners, the frequency being kept constant. For a control of the fields within 1%, the loop gain has been approximately set to 100. In practice, the smallest field unbalance we have achieved was determined by the cable and field detector's unbalance.

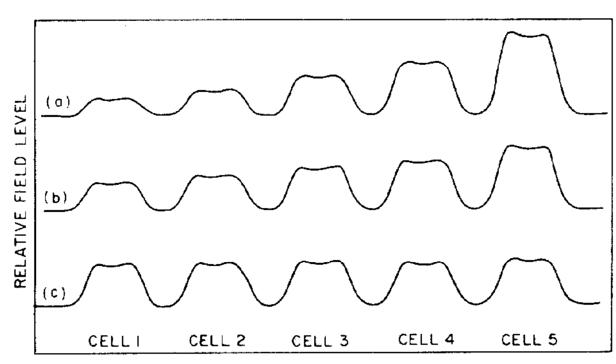


Fig. 3. Measured field profile of five cavity cells as a function of the tuners' insertion. (a) Tuner #1 inserted 10 mm, tuner #5 withdrawn 10 mm. (b) Tuner #1 inserted 5 mm, tuner #5 withdrawn 5 mm. (c) Flat field.

## 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 ±1° over several hours of operation, and over a power range of 30 dB. Figure 4 is a diagram of the phase stabilization


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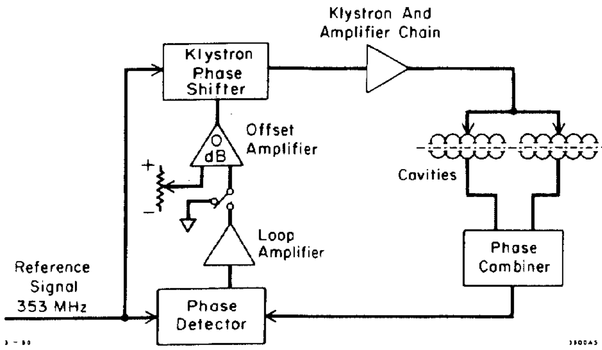


Fig. 4. Circuit of the phase control loop.

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 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. 4) without altering the phase of the total gap voltage.

Figure 5 is a layout of the phase detector scheme.[1] 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 source of ±15 mA for ±180° or transformed into an equivalent voltage source of ±5V. The dotted part of Fig. 5 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.

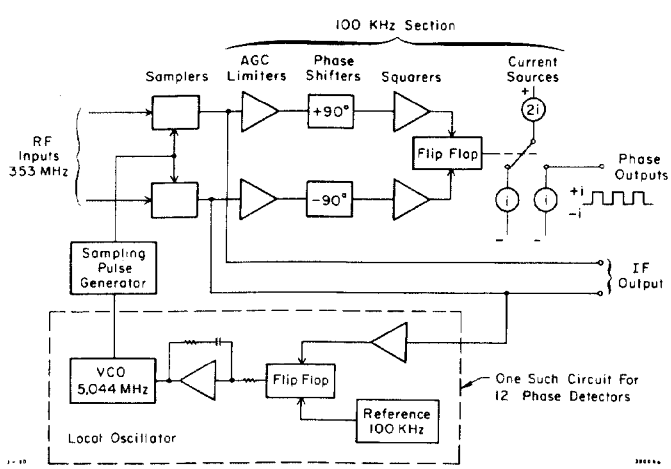


Fig. 5. Principle of the phase detectors.

* Andrew type LDF 4-50, with a phase temperature coefficient of approximately 7 ppm/°F.

Test results: A production of 40 phase detectors was evaluated. Our criterion of ±1° 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.

Electronic phase-shifters are used as control element in the phase stabilization loop, as a nulling device for each phase detector and to set the reference phase for each RF station. Depending on the use, these phase-shifters are required to have a linear response of phase versus control voltage, a phase range of up to 540° or stability of phase with temperature changes.

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 diodes (1N5461) as control elements (Fig. 6). A detailed description of the module can be found in PEP-NOTE-283.[2]

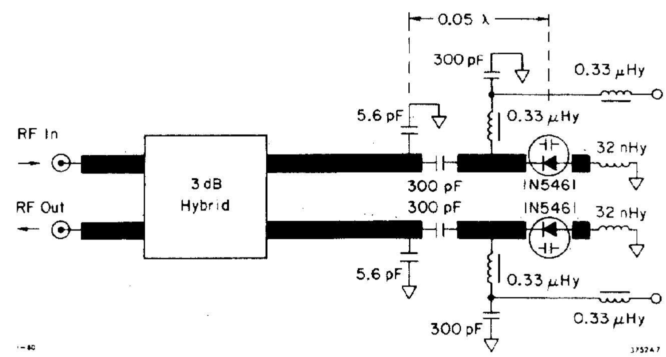


Fig. 6. 180° phase shifter module.

The typical phase-shift and insertion loss as a function of the control voltage for a 180° phase-shifter module is shown in Fig. 7. The temperature coefficient was measured to be 0.04 degrees of phase/°C.

The choice of a bandwidth for the phase control 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.

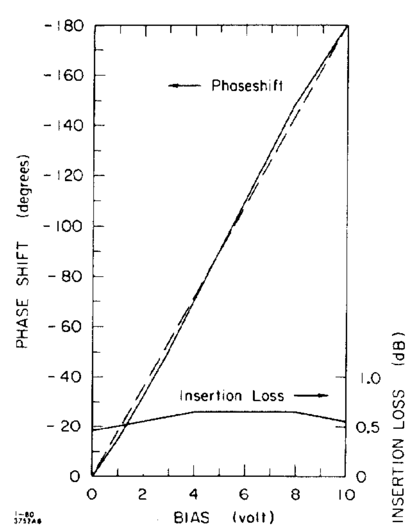


Fig. 7. Phase-shifter response and insertion loss as a function of bias voltage.

## 3. Amplitude Stabilization

A block diagram of the gap voltage feedback loop is shown in Fig. 8. The unity gain bandwidth of the amplitude loop was set to below 1 kHz to assure that the loop be insensitive to synchrotron oscillation.

The RF amplitude detector uses hot carrier diodes as peak detectors (Fig. 9). 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 compensates for 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%. An RF power of 1 watt is required to produce 10V output. The detector diodes are placed a number of half wavelength away from the sampling loop to increase their absolute accuracy. At any half wavelength point the RF voltage is largely independent of standing waves and directly related to the source voltage at the sampling loop. For this reason, all components connecting the detector to the sampling loop were made with a specified electrical length, including the approximate 100 feet length of coaxial cable (Andrew type LDF 4-50). 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 ±1.5%. Taking the other sources of error into account the accuracy with which the field level can be measured is ±5%.

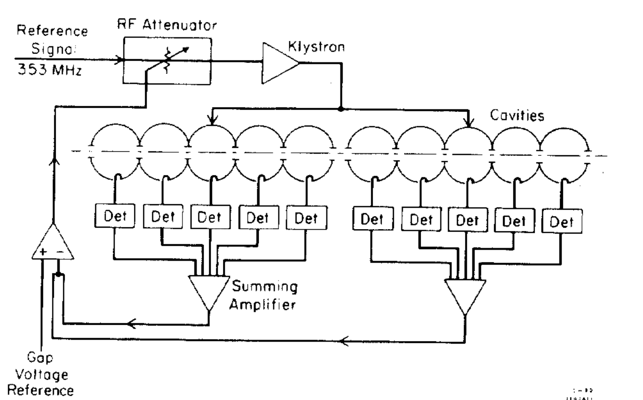


Fig. 8. Block diagram of the amplitude feedback loop.

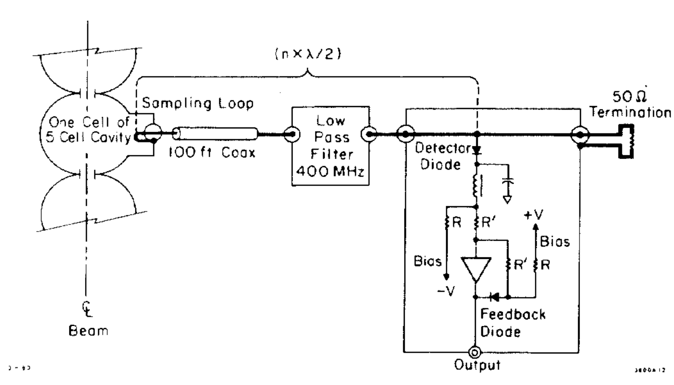


Fig. 9. RF amplitude detector circuit diagram.

A variable attenuator is used as the control element in the gap voltage feedback loop. The microstrip design is shown in Fig. 10. Two PIN diodes (HP 5082-3040) vary their individual resistance to RF from >10 kΩ to 1Ω for bias currents of zero to 50 mA. For zero bias all the input power is divided by the 3 dB hybrid (Anaren type 1S0260-3) and transmitted past the PIN diodes to be absorbed in the 50Ω terminations.

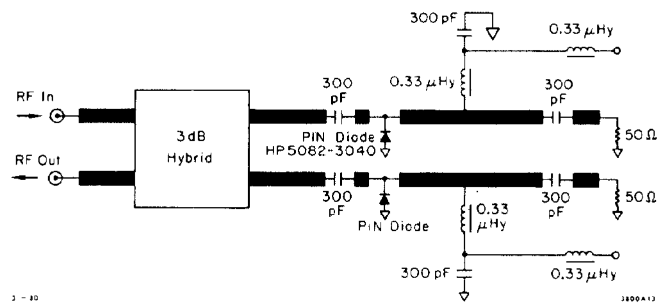


Fig. 10. Variable attenuator circuit diagram.

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. 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 with an insertion loss of about 0.8 dB.

The phase-shift variation 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 ±10°. The response of the attenuator was linearized by the addition of a nonlinear driver circuit.

## 4. Conclusions

It was considered important for the RF electronics to be accessible and easy to maintain during operation of the storage ring. All electronics components have therefore been located with the klystrons in support buildings above the storage ring tunnel. They also have been implemented in a modular format of the NIM type. 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 can only be done from the main control room via the PEP control system. A total of 12 RF stations have been operating reliably at PEP for over a year.

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

## References


1. Hewlett-Packard Vector Voltmeter 8405A Operation Manual.
2. H. Schwarz, "Linear Electronic Phase-Shifter Design," SLAC-PEP-NOTE-283, November 1979.