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ELECTRONICS FOR THE LONGITUDINAL ACTIVE DAMPING SYSTEM FOR THE CERN PS BOOSTER

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Summary and Introduction

Precisely tracking band-pass filters centred at the sixth and seventh harmonic of the revolution frequency are required[1]. During the accelerating cycle of 0.6 sec the frequency changes by a factor 2.7. The resulting tracking problem is solved by active two-path filters, where the centre frequency is governed by the frequency of a pair of sinusoidal signals in quadrature, which are generated from the accelerating RF frequency (fifth harmonic) by means of a phase-locked loop and a loop-controlled phase shifter. The phase change caused by the large frequency sweep (6 or 7 MHz) in conjunction with the delay in the feedback loop (cables, etc.) is compensated by a digital system, which computes the required phase advance from the value of the RF frequency and controls digitally the phase shift of the two-path filters. A low-frequency quadrature VCO is made to track the synchrotron frequency or harmonics hereof from analogue information about bending magnet field (momentum) and RF voltage. This quadrature pair ensures tracking of single sideband filters which permit each individual mode sideband to be examined throughout the cycle. A drive system can, by means of a similar VCO, generate any desired mode sideband, and thus excite any given mode.

Active Damping System

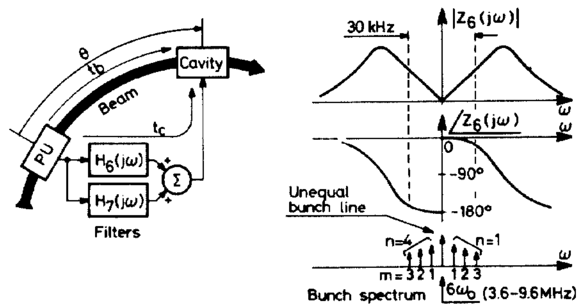

Fig. 1 Transfer functions and equivalent impedance

A longitudinal pick-up signal is passed through two band-pass filters and added to the RF signal driving the accelerating cavity (Fig. 1). This feedback loop represents an artificial coupling impedance, which must have a positive real part for frequencies slightly above the sixth harmonic and a negative real part below it, to provide damping. With a similar impedance around the seventh harmonic added, damping of the four coupled-bunch modes n = 1 to 4 with within-bunch mode numbers m = 1 to 3 is obtained[1]. The transfer functions H_6 and H_7 required to obtain this impedance depend on the feedback path delay t_c and the pick-up to cavity travelling time for the beam t_b:

$$t_b = T_0 \times \theta/2\pi = \theta/\omega_0 \ , \qquad (1)$$

where ω_0 is the revolution frequency. The impedance Z is defined by the ratio between cavity voltage and beam current at cavity:

$$\begin{aligned} I_{b,cav}(j\omega) &= I_{b,pu}(j\omega)\exp(-j\omega t_b) \\ V_{cav}(j\omega) &= H(j\omega)\, I_{b,pu}(j\omega)\exp(-j\omega t_c) \qquad (2) \\ Z(j\omega) &= V_{cav}(j\omega)/I_{b,cav}(j\omega) \end{aligned}$$

so the required transfer function becomes

$$H(j\omega) = Z(j\omega)\exp\left[j\omega(t_c - t_b)\right] = Z(j\omega)\exp(j\psi) \ . \qquad (3)$$

For frequencies near the sixth and seventh harmonics, $\omega = k\omega_0$, k = 6, 7:

$$\psi = k\omega_0 t_c - k\theta \ , \qquad (4)$$

so the phase of H_k with respect to Z_k must be offset $k\theta$ and advanced proportional to the revolution frequency.

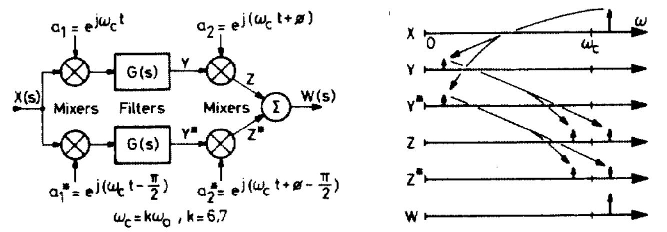

Fig. 2 Two-path active filter with phase-shift control

The transfer functions H_6,H_7 are realized by two-path filters[2], whose centre frequencies are controlled by the frequency of a pair of sinusoidal signals in quadrature (90° relative phase shift). The use of two paths eliminates the undesired sideband at the output of the second mixer (Fig. 2). Only frequencies present at the input will be present at the output, so the system behaves as a linear filter. Figure 3 shows how the two-path filter transfer function H(s) is related to G(s), the fixed filters between the mixers. A low-pass to band-pass transformation takes place; the G(s) pole-zero cluster becomes two clusters around $\pm j\omega_c$. As $|G(s-j\omega_c)| \gg |G(s+j\omega_c)|$ for $s \simeq j\omega_c$, the phase ϕ of the output mixer quadrature pair (a_2, a_2^*) can be used to control the phase shift of the two-path filter in the band-pass region. The variable phase of H(s) shows up in the pole-zero plot as a number of real zeros, whose positions depend on ϕ. The purpose of the zero at the centre frequency $j\omega_c$ is twofold. Firstly the 180° jump in phase is obtained, and secondly the unequal bunch line is suppressed by the notch in the amplitude response.

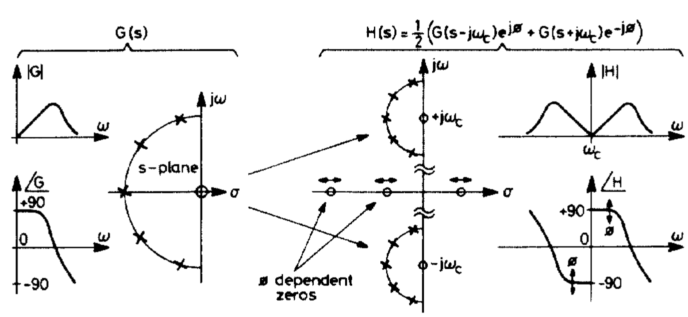

Fig. 3 Low-pass to band-pass transformation

The required phase advance could be obtained by generating the input mixer quadrature pair (a_1, a_1^*) from the output pair (a_2, a_2^*) through cables of length t_c, so (a_2, a_2^*) would have a leading phase $\phi = k\omega_0 t_c$. However, for $t_c \simeq 1.2$ µsec it would require 4 cables, each approximately 250 m long, so this solution was abandoned. Instead a digital phase control in steps of 90° is used. A 45° error in phase is acceptable as it only reduces the damping effect by $\sqrt{2}$. For 90° steps, the required (a_2, a_2^*) has a simple relation to (a_1, a_1^*) (Fig. 4). A simpler realization is, furthermore, obtained by commutating the low-frequency inputs (y, y^*) to the output mixers and leaving (a_2, a_2^*) fixed, (a_2, a_2^*) = (a_1, a_1^*).

ø	a_2	a_2^*	W
0	a_1	a_1^*	$a_1 Y + a_1^* Y^*$
+90	$-a_1^*$	a_1	$a_1 Y^* + a_1^*(-Y)$
+180	$-a_1$	$-a_1^*$	$a_1(-Y) + a_1^*(-Y^*)$
+270	a_1^*	$-a_1$	$a_1(-Y^*) + a_1^* Y$

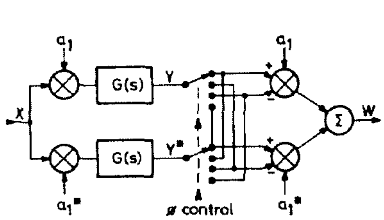

Fig. 4 Two-path filter with digital phase control

The two bits controlling the four-position switch are generated by a counter, which repetitively counts the accelerating RF frequency during a fixed count time, which is set to match the required phase advance rate (Fig. 5). The preset value of the counter prior to each counting period gives the required phase offset $k\theta$. The counter overflows each time the required phase advance exceeds 2π. The count result is stored in a 2-bit buffer register while the next counting takes place.

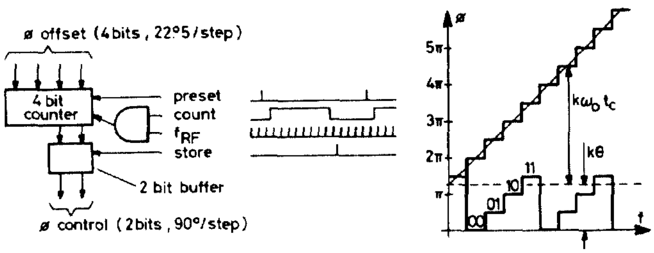

Fig. 5 Digital phase advance control

The quadrature pairs are generated by the quadrature revolution frequency generator[2,3] (Fig. 6). The frequency of a VCO divided by k is phase-locked to the accelerating frequency divided by h = 5. As the frequency of the two inputs to the phase discriminator are equal $f_k/k = f_{RF}/h$, the output frequency becomes $f_k = kf_{RF}/h$. The VCO (varicap-inductor) gives a sinusoidal output of high spectral purity and low distortion. The amplitude is kept constant by an AVC loop. The VCO output is passed through a voltage-controlled phase shifter (varicaps) to produce the quadrature component. The phase of the two outputs are compared in a phase discriminator, which controls the phase shifter to give 90° phase shift independent of frequency. A relative vector error of less than 1% has been achieved ($\Delta A < 0.1$ dB, $\Delta\phi < 0\overset{\circ}{.}6$), so a suppression better than 40 dB of the undesired sideband in the two-path filters can be obtained.

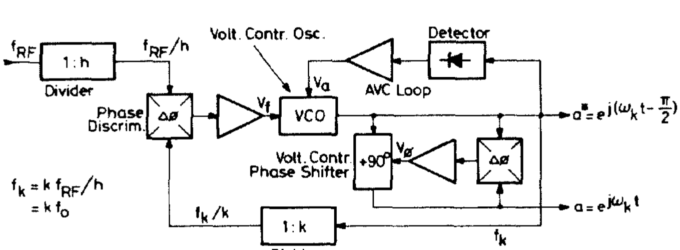

Fig. 6 Quadrature revolution frequency generator

A pick-up AVC[2] keeps the peak value of the bunch signal at a fixed level to avoid saturation of the input mixers in the filters (Fig. 7). The outputs of the filters are added to the RF signal driving the accelerating cavity. Although the impedance of the cavity at the sixth and seventh harmonics is 40 to 50 dB below its resonant impedance (fifth harmonic), sufficient voltage (50-100 V) can be obtained to get damping rates several times the highest growth rates[1]. The maximum gain is determined by noise and maximum available voltage, as the noise must not saturate the system. The fact that the accelerating cavity was used for feedback has greatly reduced the cost, as only low-level electronics has been built. The cavity has also influenced the choice of harmonics[1]. The phase offset of -90°, caused by the capacitive cavity impedance at the feedback frequencies, is compensated by the phase-advance control previously described.

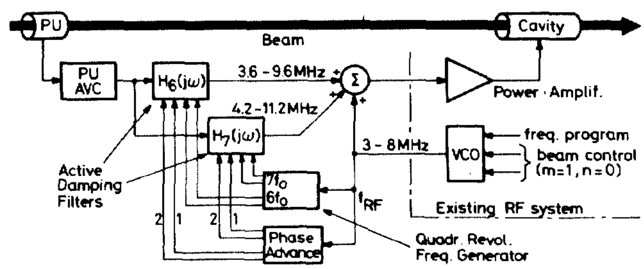

Fig. 7 Active damping system (per ring)

Mode Analysis System

Six mode sidebands (Fig. 1) are treated together in each active damping filter. Single sideband filters[2,3] with narrow bandwidth permit the examination of these individually (Fig. 8). This is important for growth and damping rate measurements with or without the damping on[1].

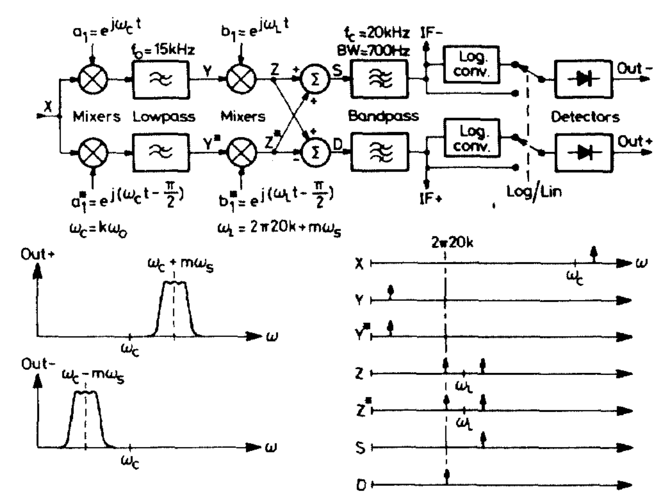

Fig. 8 Single sideband filters, mode analyser

The mixing processes are similar to the two-path filters except that the output mixers are fed by a low-frequency quadrature pair, whose frequency is related to to the synchrotron frequency as follows: f_L = 20 kHz + mf_s. The desired sideband will then appear at the output as a fixed frequency, 20 kHz. The lower sideband comes out as 20 kHz in the sum output and the upper sideband as 20 kHz in the difference output. Fixed 20 kHz band-pass filters then eliminate the undesired sidebands.

The amplitude can be observed at the output of a detector, which can operate in either linear or logarithmic mode. Sideband amplitudes from approximately 30-95 dB below the main RF line can be observed.

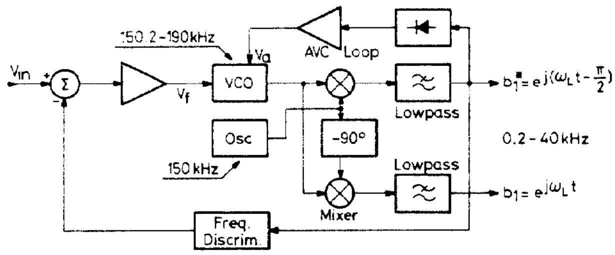

Fig. 9 Quadrature VCO, 0.2-40 kHz

The low-frequency quadrature pair is generated by a quadrature VCO [2] (Fig. 9). A fixed frequency quadrature pair is mixed with a sinusoidal signal from a VCO. A quadrature pair of variable frequency is produced by the difference mixing product. The sum frequency is suppressed by low-pass filters. A frequency discriminator stabilizes the output frequency and linearizes the voltage-to-frequency transfer function. An AVC loop keeps the output amplitude constant. The quadrature VCO is controlled by the synchrotron frequency programmer[2] to oscillate at 20 kHz + mf_s (Fig. 10). A voltage proportional to the synchrotron frequency is produced as a non-linear function of I_B (bending magnet current ∝ momentum) multiplied by the square root of the cavity voltage. Higher harmonics can be selected by a switch, and a voltage equivalent to 20 kHz is added.

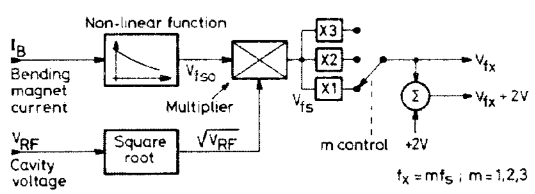

Fig. 10 Synchrotron frequency programmer

The complete mode analysis system (Fig. 11) provides parallel observation of all four coupled-bunch modes n = 1 to 4 with switch selection of the within-bunch number m = 1 to 3. Octupole modes (m = 4) have been observed on an experimental basis. A multiplexing system permits the use of the system with any of the four Booster rings. The second quadrature VCO oscillates at mf_s and is used in the drive system described below.

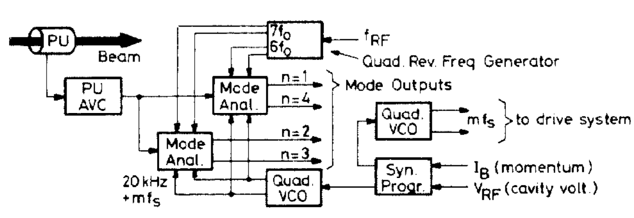

Fig. 11 Mode analysis system

Drive System, RF Knock-out

To take full advantage of the mode analysis system, it is convenient to be able to excite a single mode. This is done by driving it at its corresponding frequency $k\omega_0 \pm m\omega_s$. This sideband is generated by adding a low-frequency ($m\omega_s$) quadrature pair to the inputs of the output mixers of the active damping filters (Fig. 12). Upper or lower sidebands are selected by the sign of the c_1-c_1^* phase.

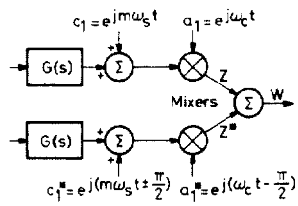

Fig. 12 Drive system

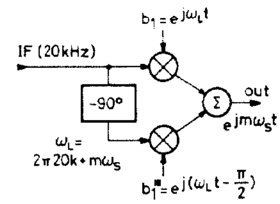

Fig. 13 Restoring drive frequency from IF.

RF knock-out consists of measuring amplitude and phase response of the beam in the neighbourhood of a mode sideband resonance. The mode analysis and drive systems are used to do this. To obtain the phase, the 20 kHz IF (intermediate frequency) of the mode analyser has to be mixed back to the driving frequency (Fig. 13).

Originally, a voltage ramp was applied to the quadrature VCO to sweep across the desired frequency range. This method was abandoned because the constant drive level was too high when the frequency was inside the band of incoherent synchrotron frequencies, where the amplitude response is strong, and too low when outside, where the amplitude response is weak. Also the bandwidth of the mode analyser (= 700 Hz) was too wide compared with the desired frequency resolution, $\Delta f_r = \sqrt{df/dt}$ = 20-40 Hz, so the signal-to-noise ratio was lower than necessary.

In the second and more successful scheme the quadrature VCO was phase-locked to an HP network analyser, which was used to measure phase and amplitude response. The frequency was changed manually and the drive level readjusted for each frequency. A Tektronix DPO (digital processing oscilloscope) was used to collect the data and convert them to a stability diagram[1]. Averaging over several cycles reduced the signal-to-noise ratio. The typical drive levels used correspond to 0.1 RF degrees centre-of-mass motion for the dipole mode.

For fixed frequency drive the frequency resolution is $\Delta f_r = 1/t_{exc}$, where t_{exc} is the excitation time. Beam dynamics considerations[4] show that the maximum drive level has to be scaled as $A_{exc} \propto (\Delta f_r)^2$. By taking into account that a bad signal-to-noise ratio has to be improved by averaging, the total measurement time scales as $t_m \propto \Delta f_{det}/(\Delta f_r)^6$, where Δf_{det} is the detector bandwidth, $\Delta f_{det} \geq \Delta f_r$. A proper choice of drive level and excitation time (resolution) is therefore crucial for a good result.

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