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

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\(^2\). The transfer functions \(H_6\) and \(H_7\) required to obtain this impedance depend on the feedback path delay \(\tau_c\) and the pick-up to cavity traveling time for the beam \(\tau_b\):

\[
\tau_c = T_0 \times \frac{6}{2\pi} = \frac{\theta}{\omega_0},
\]

where \(\omega_0\) is the revolution frequency. The impedance \(Z\) is defined by the ratio between cavity voltage and beam current at cavity:

\[
I_{ave}(j\omega) = I_{b,cav}(j\omega) \exp\left(-j\omega \tau_b\right)
\]

\[
V_{cav}(j\omega) = H(j\omega) I_{b,cav}(j\omega) \exp\left(-j\omega \tau_c\right)
\]

\[
Z(j\omega) = \frac{V_{cav}(j\omega)}{I_{ave}(j\omega)} = H(j\omega) \exp\left(j\omega \tau_c\right)
\]

so the required transfer function becomes

\[
H(j\omega) = Z(j\omega) \exp\left[j\omega (\tau_c - \tau_b)\right] = Z(j\omega) \exp\left(\psi j\omega\right).
\]

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

\[
\psi = k\omega_0 \tau_c - k\theta,
\]

so the phase of \(H_6\) with respect to \(Z_6\) must be offset \(k\theta\) and advanced proportional to the revolution frequency.

The transfer functions \(H_6, H_7\) are realized by two-path filters\(^5\), 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 \(j\omega_0\). As \(\left|G(s-j\omega_0)\right| > \left|G(s+j\omega_0)\right|\) for \(s = j\omega_0\), the phase \(\phi\) of the output mixer quadrature pair \((a_1, a_2)\) 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 \(\phi\). The purpose of the zero at the centre frequency \(j\omega_0\) 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. 1 Transfer functions and equivalent impedance

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

Fig. 3 Low-pass to band-pass transformation
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
\((a_2, a_3)\) has a simple relation to \((a_1, a_3)\) (Fig. 4). A
simpler realization is, furthermore, obtained by commu-
tating the low-frequency inputs \((y, y^*)\) to the output
mixers and leaving \((a_2, a_3)\) fixed, \((a_2, a_3) = (a_1, a_3)\).

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 \(k\phi\). The
counter overflows each time the required phase advance
exceeds \(2\pi\). 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 quadra-
ture revolution frequency generator (Fig. 6). The
frequency of a VCO divided by \(k\) is phase-locked to the
accelerating frequency divided by \(h = 5\). As the fre-
quency of the two inputs to the phase discriminator are
equal \(\omega_k/k = \omega_{SR}/h\), the output frequency becomes
\(\omega_k = \omega_{SR}/h\). 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 out-
put 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 dis-
criminator, 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, \(\Delta \phi < 0.5°\)), 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 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 90 dB below
its resonant impedance (fifth harmonic), sufficient volt-
age (50-100 V) can be obtained to get damping rates
several times the highest growth rates. The maximum
gain is determined by noise and maximum available volt-
age, so 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 electron-
sics has been built. The cavity has also influenced
the choice of harmonics. The phase offset of -90°,
caused by the capacitive cavity impedance at the feed-
back 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 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.

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 the synchrotron frequency as follows: \(f_k = 20\) kHz +
+ \(m\Delta f\). The desired sideband will then appear at the
output as a fixed frequency, 20 kHz. The lower sideband
comes out at 20 kHz in the sum output and the upper side-
band at 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.

The low-frequency quadrature pair is generated by a quadrature VCO (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 to oscillate at 20 kHz ± 10 kHz. A voltage proportional to the synchrotron frequency is produced as a non-linear function of \( I_B \) (bending magnet current or 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.

The complete mode analysis system (Fig. 11) provides parallel observation of all four coupled-bunch modes \( n = 1 \) 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 \( f_m \) and is used in the drive system described below.

### 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.

### References

1. F. Pedersen and F. Sacherer, Theory and performance of the longitudinal active damping system for the CERN PS Booster, these proceedings.