LBL-34405
CBP Note-030
PEP-II AP Note-37-93
PEP-II EE Note-05-93

ELECTRONIC SYSTEMS FOR TRANSVERSE COUPLED-BUNCH FEEDBACK IN THE ADVANCED LIGHT SOURCE (ALS)*

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Submitted to the Beam Instrumentation Workshop, Santa Fe, NM
October 20 - 23, 1993

* This work was supported by the Director, Office of Energy Research, Office of Basic Energy Sciences, Materials Sciences Division, of the U. S. Department of Energy, under Contract No. DE-AC03-76SF00098.

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ABSTRACT

In order to effectively control a large number of transverse coupled-bunch modes in the LBL Advanced Light Source (ALS) storage ring, a broad-band, bunch-by-bunch feedback system has been designed [1], and is beginning to undergo testing and commissioning. This paper addresses, in some detail, the major electronic components of the feedback system. In particular, the components described include: broad-band microwave position detection receivers, closed orbit offset signal rejection circuitry, and baseband quadrature processing circuitry.

INTRODUCTION

The LBL Advanced Light Source (ALS) is a third generation 1.5 GeV electron storage ring for producing synchrotron radiation in the 0.5 - 1000 eV range [2]. Because of the high average beam current in the ring (400 mA), active feedback systems for damping longitudinal and transverse coupled-bunch instabilities are required. The storage ring is designed to accommodate a large number of bunches, up to 328 in buckets separated by 2 ns (500 MHz RF). As a result, a broad spectrum of transverse coupled-bunch modes are driven by the higher-order transverse modes of the RF cavities and the transverse resistive-wall impedance [3]. In order to effectively damp and control growth of these modes, a 250 MHz bandwidth, bunch-by-bunch feedback system has been designed [1], and is presently undergoing fabrication, testing, and commissioning. In this paper, the engineering details of the transverse feedback system electronics are presented and discussed.

SYSTEM OVERVIEW

The overall transverse feedback system concept is shown in Figure 1. The system utilizes two pickup stations, each of which is used to detect both horizontal and vertical beam moment $(I\Delta X_{\perp})$. By summing moment signals from the two pickups in proper proportions, kick signals that are exactly in quadrature with horizontal and vertical beam position at the kickers can be produced. Because the relative contributions of the two pickups is adjustable, this technique allows for arbitrary kicker placement and can accommodate a wide range of betatron tunes.

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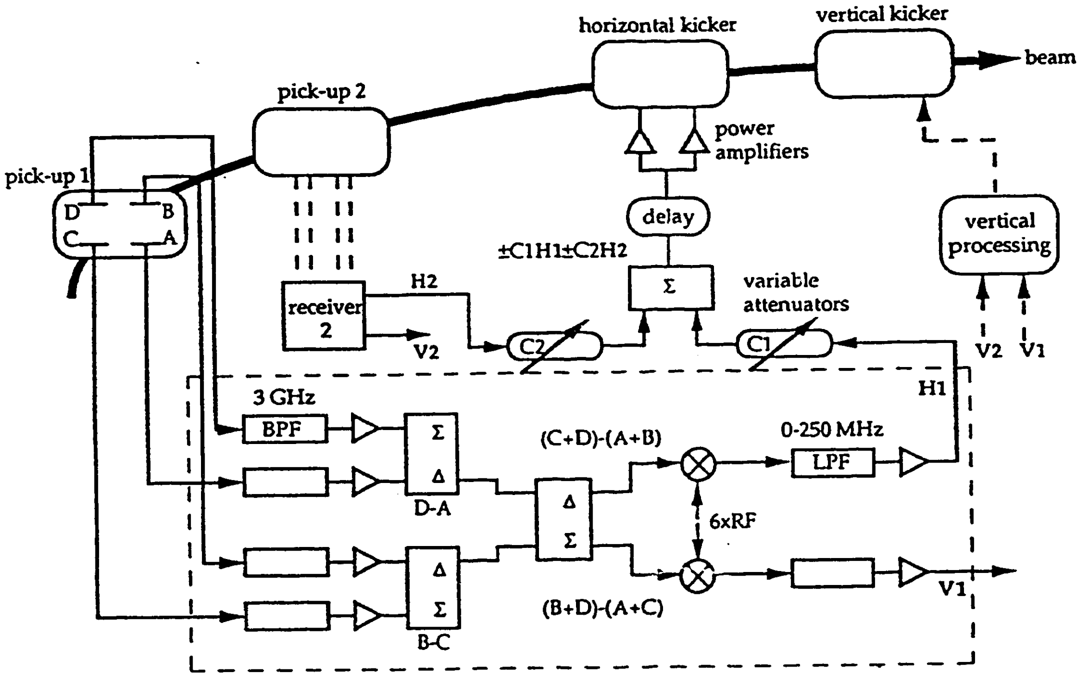

Figure 1 Feedback System Overview

The pickups consist of standard LEP buttons modified to suppress high frequency resonances [4]. Separate horizontal and vertical stripline kickers are used to obtain maximum efficiency by optimizing the electrode coverage factor. The electronics systems consist of two microwave receivers for detecting horizontal and vertical beam moment, a system (shown schematically as two variable attenuators) for mixing the signals from the two pickup stations, and four broad-band power amplifiers for driving each electrode of the two kickers. Finally, coaxial delay lines are used to make moment signals picked up from a given bunch coincident with the arrival of that same bunch at the kickers.

The 500 MHz bunch rate in the ALS dictates a minimum 250 MHz base-bandwidth for a bunch-by-bunch feedback system. As long as this bandwidth criterion is met, moment detection as well as kicking may be performed using any harmonic of the 500 MHz bunch rate as a carrier. Because of the greater efficiency advantages of low-frequency transverse kickers, the baseband range (~ 150 kHz - 250 MHz) is used for kicking the beam. The low-frequency end of this band is determined by the revolution frequency, 1.5 MHz, and the assumption that the fractional part of the tune will not exceed 0.9. The upper end of the passband, 250 MHz, allows for the possibility of having to damp the coupled bunch mode where bunches essentially alternate plus and minus in position as they pass a fixed point in the ring.

One drawback of kicking in the baseband is the problem of obtaining high power (150 W each) microwave amplifiers that have good linearity and phase characteristics over this relatively wide band. After an exhaustive search, it was found that the maximum cutoff frequency of readily available amplifiers that meet our specifications is 220 MHz. Fortunately, measurements of the higher order mode spectrum of the ALS RF cavities indicate that transverse modes with

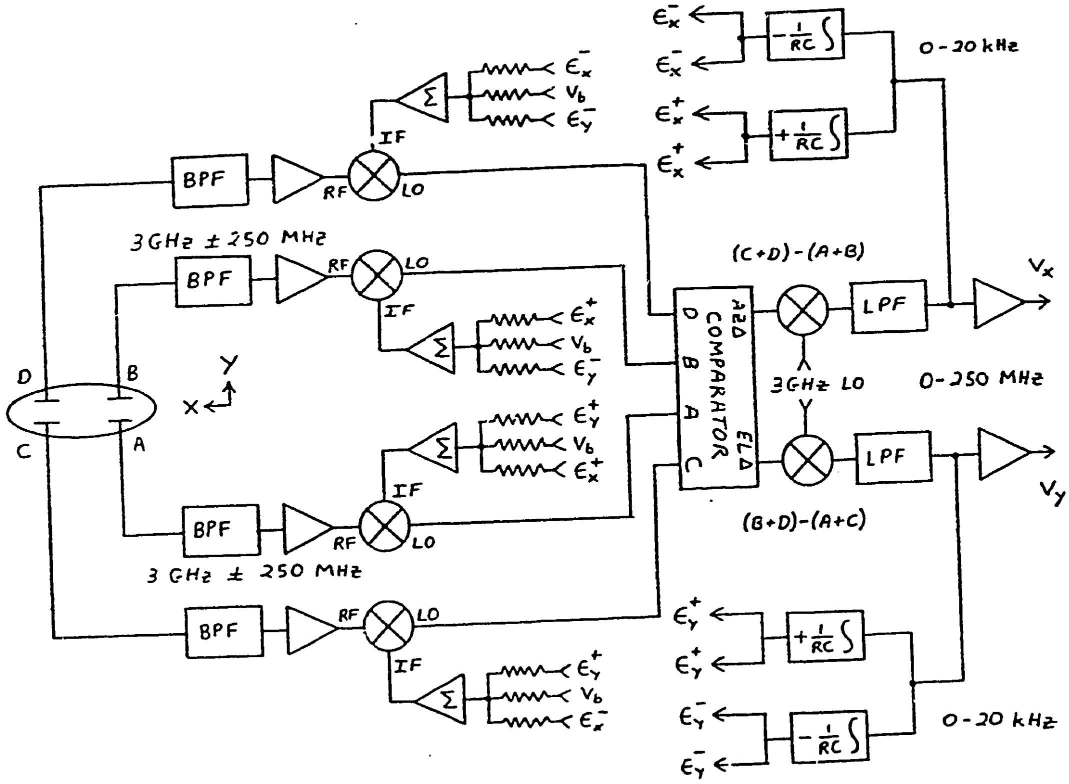

Figure 2 Front-End Receiver

impedances large enough to drive instabilities are not present in the band gaps n x 250 MHz ± 30 MHz where n is an odd integer [5]. Therefore, the 220 MHz amplifiers are expected to be sufficient for the ALS system. The 150 W amplifiers (300 W per kicker) will provide transverse kicks per turn ranging from 2.3 kV at 100 kHz to 1.6 kV at 220 MHz. At the nominal betatron tunes, these voltages are sufficient to control any expected transverse coupled-bunch motion. In contrast to the kicker band of operation, detection is performed at 3 GHz where the pickups are most sensitive. As indicated in Figure 1, the front-end receivers detect beam moment at this frequency and subsequently demodulate the moment signals to baseband for driving the kickers.

FRONT-END RECEIVERS

The 3 GHz front-end receiver design is shown in more detail in Figure 2. Bunch-to-bunch position variations due to coupled-bunch instabilities result in amplitude modulation of the 3 GHz harmonic of the beam current (carrier) detected by the button pickups. The pickup signals are routed to the receivers via 100 ft. lengths of phase matched cables. In addition, the receiver inputs employ adjustable delay lines (not shown) for further trimming of the phase match between channels. The button signals are band-limited with 3 GHz ± 250 MHz bandpass filters (BPF's) allowing for double sideband AM detection. The BPF's also prevent intense spikes from individual bunches from reaching the front-end amplifiers. Apart from the mixers after the amplifiers, whose functions will be described shortly, the remaining portions of the receivers simply sum and difference the button signals using monopulse comparators to produce x and y moment signals, V_x and V_y. The x and y signals are synchronously demodulated to baseband with a 3 GHz local oscillator locked to the storage ring RF.

The signal-to-noise performance of the receivers is determined by the pickup impedances, the cable losses from the pickups to the receivers, the front-end amplifier noise figures, and the system bandwidth. Assuming that cable losses are 6 dB and the noise figure for the front-end amplifiers is 3 dB, the signal-to-noise at the output of the high power amplifiers is given by:

$$\frac{S}{N} = \frac{(Z_\perp I_o X_\perp)^2}{16 Z_o k T_o \Delta f_s} \tag{1}$$

where:

	$Z_\perp$	=	x, y pickup impedance
	$X_\perp$	=	x, y beam displacement (amplitude)
	I_o	=	average <u>bunch to bunch</u> beam current – .5 amp
	Δf_s	=	system bandwidth – 220 MHz
	Z_o	=	50 Ω
	kT_o	=	4×10^{-21} W/Hz

It is desirable to damp the amplitude of transverse beam motion to one tenth the beam size, which for ALS, is about 200 µm rms in x and 40 µm rms in y at the pickup and kicker locations. For the button pickups, $Z_{\perp x} = 2$ Ω/cm and $Z_{\perp y} = 0.5$ Ω/cm. These give $S/N_x = 38$ dB and $S/N_y = 12$ dB for 20 µm and 4 µm amplitudes in the x and y directions respectively.

Because broadband RF power is expensive, it is important that the feedback system not respond to closed orbit offsets that can easily be corrected with steering magnets. In order to reject signals due to closed orbit offsets, the receivers employ feedback loops that eliminate beam-moment signals due to slowly varying beam position by changing the gains of each button signal so that on average, the beam always appears to be centered in the button array. Referring to Figure 2, beam-moment signals at the outputs of the x and y low pass filters are integrated to produce error signals that drive the DC-coupled IF ports of the mixers in each button signal path. The mixers are configured in "attenuator" mode so that with the proper bias, V_b, the signal attenuation from the RF to the LO port of each mixer is linearly proportional to the error signal at the IF port. This produces the gain control over each button signal required to zero the moment signals at the low pass filter outputs. The RC time constant of the integrators is set so that only moment signals caused by beam position variations with frequencies less than approximately 20 kHz are zeroed. These would include signals resulting from static orbit offset, transverse beam motion due to magnet power supply ripple, and if present, transverse motion at the synchrotron frequency. Beam moment signals resulting from position variations faster than 20 kHz (i.e., betatron oscillations) are passed undisturbed to the kickers. An added benefit of the offset feedback loops is that they also compensate for phase and amplitude imbalances in the front-end RF components and the comparator.

The offset feedback loops are a somewhat novel feature of the receivers so they will be presented here in a more quantitative manner. In order to make the discussion more tractable, the simpler two-button case shown in Figure 3 is analyzed. The filtering and demodulating processes have been lumped into single blocks which precede the variable gain elements and the differencing element. The

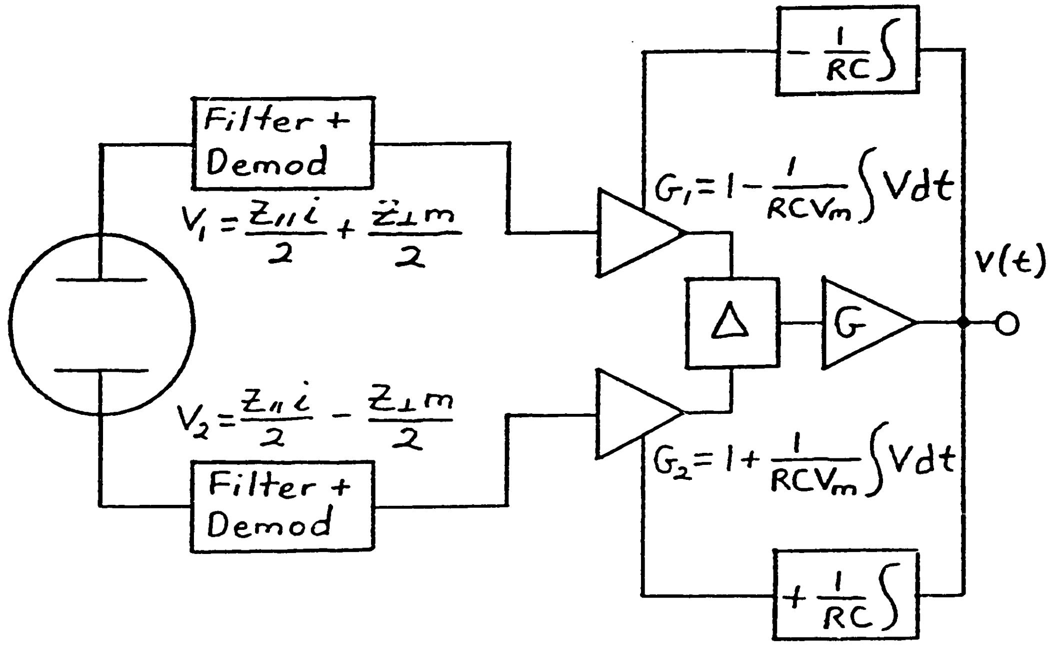

Figure 3 Circuit for Offset Feedback Loop Analysis

total gain of the receiver, G, is represented by a block after the differencing element. This rearrangement of the receiver components allows the system to be more clearly analyzed in terms of baseband signals and is permissible as long as the bandwidths of the signals driving the variable gains are small compared to the system bandwidth.

The voltages, $v_1(t)$ and $v_2(t)$, are the baseband (filtered and demodulated) signals from each button pickup. Each of these signals consists of position-dependent and independent components, $Z_{\perp}m(t)/2$ and $Z_{\parallel}\, i(t)/2$. Here, $i(t)$ and $m(t)$ are the detected filtered and demodulated beam current and moment signals. The impedances, $Z_{\parallel}$ and $Z_{\perp}$, are the longitudinal and transverse pickup impedances of the pickup pair which are treated as simple multiplicative constants because their frequency responses are flat over the 3 GHz ± 250 MHz band. The gains, G_1 and G_2, are linearly proportional to error signals derived from the integral of the output signal. Note that V_m is the maximum output of the integrators.

Qualitatively, the action of the feedback loops is seen by considering, a stationary beam displaced towards the upper button pickup. In this case, G_1 is decreased and G_2 is increased by the same amount so that $v(t) = 0$ while maintaining a constant system gain. The starting point for analyzing the frequency response of the system is the circuit equation for the feedback loops:

$$\frac{Z_{\parallel}\, i(t)}{RCV_m}\int v(t)\, dt - Z_{\perp}\, m(t) - \frac{v(t)}{G}. \qquad (2)$$

Equation (1) is easily transformed into a differential equation for $v(t)$:

$$\frac{dv(t)}{dt} + \left[\frac{GZ_{\parallel}}{RCV_m}\, i(t) - \frac{1}{i(t)}\frac{di(t)}{dt}\right] v(t) = GZ_{\perp}\, i(t)\, \frac{d}{dt}\left(\frac{m(t)}{i(t)}\right). \qquad (3)$$

Equation (3) is a first order linear differential equation for $v(t)$. However, the equation coefficients are not constant making it difficult to analyze the system response using the usual transfer function methods. This is to be expected because the system is not time invariant due to the gains, G_1 and G_2, which vary with time. Even though transfer function methods are difficult to apply, equation (3) can still be solved in the time domain by elementary methods:

$$v(t) = GZ_{\perp}\, i(t)\; e^{-\frac{GZ_{\parallel}}{RCV_m}\int i(t)\,dt} \int \left(\frac{m(t)}{i(t)}\right)' e^{\frac{GZ_{\parallel}}{RCV_m}\int i(t)\,dt}\, dt. \qquad (4)$$

Although the above solution is exact, it is still difficult to obtain frequency response information from it. In order to proceed further, we use the approximation that the loop bandwidth is negligibly small compared to the revolution frequency (20 kHz vs. 1.5 MHz). In this case, (4) can be written:

$$v(t) = GZ_{\perp}\, i(t)\, e^{-t/\tau} \int \chi'(t)\, e^{t/\tau}\, dt \qquad (5)$$

where: $\chi(t) = m(t)/i(t)$ (closely related to beam position)
$\tau = RCV_m/GZ_{\parallel}I_{av}$
I_{av} = average current in the storage ring

Expression (5) can be written in the equivalent form:

$$v(t) \approx GZ_{\perp}\, i(t) \int_0^t \chi'(\lambda)\, e^{-(t-\lambda)/\tau}\, d\lambda \qquad (6)$$

so that:

$$v(t) = GZ_{\perp} i(t)\, L^{-1}\left\{\chi(s)\left[\frac{s}{s+1/\tau}\right]\right\} \qquad (7)$$

where L^{-1} indicates the inverse Laplace transform.

Expression (7) is the desired result because the frequency response of the system appears explicitly in the square bracketed term. The response of the system to χ is clearly that of a simple first order high pass filter with cutoff $\omega_c = 1/\tau$. It should be noted that ω_c and τ depend on I_{av}. This results because the effective gain of the feedback depends on I_{av}. The strategy for setting ω_c is to set RC so that $f_c \approx 20$ kHz at full beam current (0.4 amp). In this case, beam position signals at 10 kHz that could cause synchrobetatron coupling are prevented from reaching the power amplifiers and kickers. At lower currents, the rejection of these signals is reduced but so are the demands on the power amplifiers so that some expenditure of power on the synchrotron signals should not pose any problems. In all cases, static orbit offset and front-end component imbalances are corrected by the offset feedback loops. At frequencies above f_c, the output approaches, $v(t) \approx$

$GZ_\perp m(t)$ so that moment signals due to betatron oscillations are passed as desired.

The response of the offset feedback system, subject to the approximations above, with a general feedback element, *H(s)*, may be found by substituting *H(s)* for $1/RCs$ in equation (7):

$$v(t) = GZ_\perp i(t)\, L^{-1}\left\{\frac{\chi(s)}{1+G'\,H(s)}\right\} \tag{8}$$

where:

$$G' = \frac{GZ_\parallel I_{av}}{V_m}.$$

Using equation (8), the response and stability of the offset feedback loop with general *H(s)* can be analyzed with standard techniques. The offset feedback loop design for the ALS actually contains a first order low pass filter as well as an integrator in the feedback path in order to provide a sharper rolloff. This configuration is easily shown to be unconditionally stable using equation (8).

DUAL PICKUP QUADRATURE PROCESSING

Referring to Figure 1, the feedback system utilizes two sets of pickups with adjustable coefficients in order to produce a kick signal that is in quadrature with beam position at the kicker. The quadrature condition ensures that the area of the phase space ellipse is reduced on every turn by this proportional feedback [6]. This type of feedback is simple and efficient for power limited systems, i.e. where the maximum possible transverse kicker voltage is small compared to the maximum transverse momentum of a bunch. It can be shown that the coefficients required for quadrature processing are given by:

$$C_1 = \frac{\cos\psi_{23}}{\sqrt{\beta_1\beta_3}\,\sin\psi_{12}} \tag{9}$$

$$C_2 = \frac{-\cos\psi_{13}}{\sqrt{\beta_2\beta_3}\,\sin\psi_{12}} \tag{10}$$

where: $\beta_1, \beta_2, \beta_3 = \beta$ functions at pickup 1, pickup 2, and kicker respectively.
ψ = betatron phase advance between elements indicated by subscripts.

Referring to the right half of the block diagram of Figure 4, the coefficients can be directly applied to the moment signals with mixers configured in the attenuator mode. As indicated, the moment signals multiplied by the correct coefficients are summed to produce the quadrature kick signals, $X_k = C_{1x} V_{x1} + C_{2x} V_{x2}$ and $Y_k = C_{1y} V_{y1} + C_{2y} V_{y2}$. In this direct implementation scenario, the coefficients are calculated from equations (9) and (10) using β's and ψ's obtained from an on line lattice program.

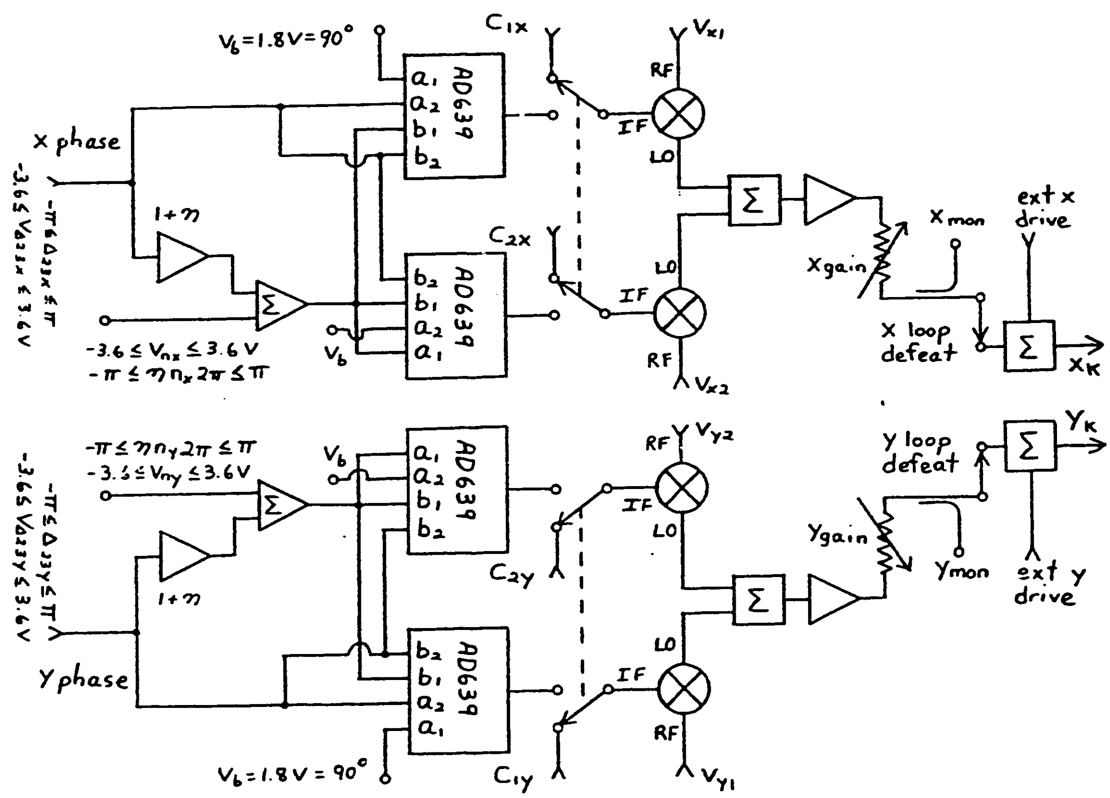

Fig. 4 Quadrature Processing Circuitry

Setting the coefficients in the direct implementation case involves a two parameter adjustment. It is also desirable to have a "single knob" adjustment which produces the correct coefficients, C_1 and C_2 in case the betatron tune is changed. Operationally, this would greatly simplify the setting of the feedback system because some observable such as damping rate or kick/position quadrature at the kicker can be optimized with a single adjustment. In order to implement a single knob scheme; the coefficient expressions (9) and (10), must be examined for the specific case of the ALS lattice. Apart from insertion devices, the lattice consists of twelve identical cells joined by straight sections. The transverse feedback system pickups are located at identical points in the unit cell (at the beginning of the straight section) and are separated by one cell. In this case, $\beta_1 = \beta_2 = \beta_p$ and expressions (9) and (10) can be written, apart from a factor of $1/\sqrt{\beta_p \beta_3}$ which is common to both, as:

$$C_1 \,\alpha\, \frac{\cos \psi_{23}}{\sin \psi_{12}} \tag{11}$$

$$C_2 \,\alpha\, \frac{-\cos \psi_{13}}{\sin \psi_{12}} \tag{12}$$

The kickers are also located in a straight section ten cells downstream from the pickup so that $\beta_3 \approx \beta_p$. Therefore, neglecting the effect of insertion devices, one would expect the ratio of phase advances, $\psi_{12}/\psi_{23} - \eta$, to be constant and in particular, to be equal to 0.1. The phase advances can now be written:

$$\psi_{23} - \Delta_{23} + n2\pi$$
$$\psi_{12} - \eta\Delta_{23} + \eta n2\pi$$
$$\psi_{13} - (1+\eta)\Delta_{23} + \eta n2\pi$$

The coefficients can now be expressed in terms of the single variable Δ_{23}, which is the principal value of the phase advance between the second pickup and the kicker:

$$C_1 \; \alpha \; \frac{\cos \Delta_{23}}{\sin(\eta\Delta_{23} + \eta n2\pi)} \tag{13}$$

$$C_2 \; \alpha \; \frac{-\cos\left[(1+\eta)\Delta_{23} + \eta n2\pi\right]}{\sin(\eta\Delta_{23} + \eta n2\pi)} \tag{14}$$

The effect of the insertion devices on the constancy of η is expected to be negligible for the ALS storage ring. Even if this were not the case, there is always a value of the single parameter, Δ_{23}, which produces kick/position quadrature at the kicker. The non-constancy of η simply results in a small variation in the gain of the feedback system when Δ_{23} is adjusted to obtain quadrature after an intentional or insertion-device-induced tune shift. If necessary, the gain variation can be simply compensated with the system gain adjustments shown in Figure 4.

The single-parameter expressions, (13) and (14), are electronically implemented with Analog Devices AD639 trigonometric function converters. The AD639 produces an output voltage given by:

$$V_{out} = A \frac{\sin(a_1 - a_2)}{\sin(b_1 - b_2)} \tag{15}$$

The angles, a_1, a_2, b_1, and b_2, are equal to the corresponding input voltages scaled by 50°/volt. The factor, A, is an arbitrary amplitude. By tracing through the circuitry and using the AD639 property given by (15), one can confirm that IF port control voltages which are proportional to the coefficients given by (13) and (14) are synthesized. The factor, η, enters into the circuit as a gain block, $1 + \eta$, and the factor, $\eta n2\pi$, enters in as a fixed bias. These two quantities are fixed during the initial setup of the feedback system and need be changed only in the event of a drastic alteration of the lattice or a change in tune between the second pickup and the kicker greater than an integer. The single knob adjustment, labeled as x phase (or y phase), corresponds to Δ_{23} and has a 360° range. Thus, setting of the system is accomplished by the simple and convenient procedure of adjusting the phase knobs until quadrature is obtained at the kickers. The degree of quadrature at the kickers is monitored with a phase detector which measures the phase of the kick signals verses the position signals obtained from the upstream ports of the stripline kickers. Finally, Figure 4 illustrates some of the more important minor features of the feedback system such as monitor points and external drive inputs for making tune and beam transfer function measurements.

SUMMARY

A feedback system for controlling transverse coupled-bunch instabilities in the LBL ALS storage ring has been designed and will soon undergo testing and commissioning. The design philosophy has been one of simplicity and user friendliness. To this end, a simple, novel technique for phasing the system in the face of changes in machine tune has been designed and described. In addition, other major features of the system such as orbit offset correction circuitry have been addressed.

ACKNOWLEDGMENTS

The authors wish to thank J. Byrd, J. Corlett, J. Fox, and many other members of the LBL/SLAC - ALS/PEP II feedback systems collaboration for many enlightening discussions and useful suggestions.

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