 IEEE Transactions on Nuclear Science, Vol. NS-32, No. 5, October 1985

CONTROL OF CAVITIES WITH HIGH BEAM LOADING

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Abstract

Beam loading of RF cavities is a subject of great concern for the design and operation of high current circular accelerators and storage rings. The steady state and transient perturbations of the accelerating voltage by the beam current lead to undesirable beam behaviour, like for instance the onset of instabilities or particle loss by lack of longitudinal acceptance. To some extent, it is possible to alleviate or even completely suppress these effects by a proper control of the RF power amplifier - cavity combination, using feedback or feedforward techniques. Several new developments in this field have taken place during the past years. For instance, the implementation of fast feedback technology, the use of cavity compensation schemes or digital filtering of signals in long delay feedback systems have resulted in considerably improved machine performance. Such techniques will be reviewed in this paper and their performance and limitations will be presented.

1. The equivalent circuit

It is customary to analyse the problem of beam loading on RF cavities with the equivalent circuit of Fig.1, [1, 2, 3, 4] which represents the RF cavity, in the vicinity of the main resonance, seen from the accelerating gap. By a proper transformation from the RF generator to the gap, the power tube can be described by the current source i_g, the generator impedance being included into the RLC circuit elements. We assume, for simplicity that the two current sources i_g and i_b (component of beam current at the RF frequency f_{RF}) flow into the same impedance. This may not be the case if there is a long line between generator and cavity or if the accelerating structure is of the travelling wave type [5].

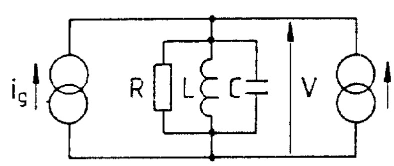

Fig.1 Equivalent circuit

Taking V (accelerating voltage) as reference, the phasor diagram of Fig. 2 represents the circuit behaviour at f_{RF}. For a single cavity (or if all cavity voltages are in phase), V and i_b are in quadrature for no acceleration ($\Phi_s = 0$); in the general case their phase difference is $\Phi_s + \pi/2$. The total current $i_t = i_g + i_b$ is related to V by the cavity admittance :

$$i_t = i_g + i_b = YV = (\frac{1}{R} + jB)\ V \qquad (1)$$

The vector i_t follows the dotted line of Fig.2 when the tune (jB) of the cavity is varied (R = constant). The quantity i_g is determined by the cavity through the relation $i_g = i_t - i_b$. Obviously $|i_g|$ is minimum when V and i_g are in phase : this is the best operating point for the power generator; there is usually a servo-tuning system designed to maintain this condition and to minimize the required RF power. In other words the tune of the cavity (at f_{RF}) is controlled by a feedback system which keeps V and i_g in phase ; as a result all the generator power (active power) is entirely converted into beam acceleration and cavity losses. However, in some cases the cavity tuning range is not large enough to provide the reactive power needed to compensate the steady beam loading. There, the RF generator must work on a non resistive load and deliver reactive power.

We have described here an equilibrium situation and must now examine the stability of the system. Consider for instance the transient regime of the cavity when i_b changes rapidly, like at injection of a prebunched beam [6], during adiabatic capture or when a fraction of the beam is fast extracted [7].

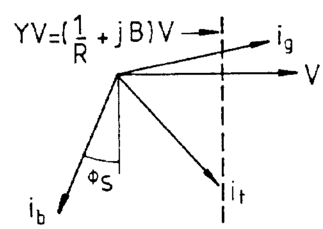

Fig. 2
Phasor diagram

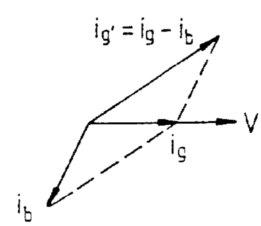

Fig. 3
Correction of beam loading transient with the power generator

The resulting variation of V must be properly damped, and in addition, it must be short enough compared to the synchrotron period T_s in order to avoid significant effects on the beam (mismatch and subsequent blow-up, or even loss of particles). If this condition is fulfilled, fundamental beam loading effect will be unimportant.

2. Peak generator power

The worst transient case is when a prebunched beam i_b is injected into an empty machine. We shall take this as an example in the following. Before injection, the servo-tuning keeps $i_t = i_g$ and V in phase. Immediately after injection, the new vector i_b destroys the equilibrium, and V changes by a large amount until the tuning loop retunes the cavity to a different value. Unless one uses very fast tuners [8] (which may lead to multiloop stability problems) [4], it will take more than a small fraction of a synchrotron period for the tuning loop to settle at its new value, the result being a strong distortion of the longitudinal phase plane.

The only way to maintain V constant during the transient phase of the tuner is to act via the RF power generator which provides a fast control of V.

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The obvious solution (Fig. 3) is to change i_g into i_g' when the beam is injected. If we make :

$$i_g' = i_g - i_b \qquad (2)$$

the total current in the cavity does not change, and, at constant tuning, V stays constant.

In the simple case of no acceleration, the amplitude of the peak current i_g' which must be delivered by the RF power tube during the transient phase of the tuner, is given by:

$$|i_g'|^2 = |i_g|^2 + |i_b|^2 \qquad (3)$$

This extra current must be delivered in a non matched load, in this simplified example. With a circulator inserted between the RF amplifier and the cavity (Fig. 4), the generator is always matched and the extra current also means extra power. Again for Φ_s=0 a similar analysis can be done in this case ; it gives the peak power P' needed during the transient phase of the tuner [9] [10]:

$$P = P_o\left[1 + \left(\frac{|V||i_b|}{4\,P_o}\right)^2\right] \qquad (4)$$

P_o is the power for no beam (matched cavity); the excess power P' - P_o is simply wasted into the load to keep V constant. One can optimize P' by selecting the best cavity impedance (R_{opt} = $2V/i_b$) and obtain the simple result:

$$P'_{opt} = 2P_o = |V||i_b|/2 \qquad (5)$$

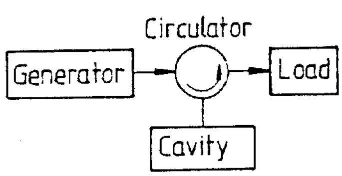

Fig.4 A circulator to match the RF power generator

Remember nevertheless that this is the worst case situation, and in certain cases, it is possible to minimize the required peak power (or peak current). In particular, by pretuning the cavity before injection, one can make the two powers, before and after injection, equal and obtain in this case $P'_{opt} = |V||i_b|/4$ (for Φ_s=0) [9]. With superconducting cavities, (usually without variable tuners) the peak power can even be reduced to $|V|\,|i_b|/8$ [9] [10]. One can also reduce the transient on i_b with multiple injections of smaller currents, or by adjusting the bunching factor of the injected beam.

In the above analysis, we assumed the filling time of the cavity to be long compared to the revolution period T_o = $1/f_o$ (but small with respect to T_s): all bunches were submitted to the same RF voltage. If this is not the case (Q < h ; Q: quality factor of the cavity, h: harmonic number) unequal filling of the ring will give a modulation of V at f_o and its multiples. The same analysis applies here : at each "batch" passage transient beam loading must be corrected to make all bunches see the same RF voltage. This effect is particularly important in large machines, not only at injection, but also at transition [11] [12]. As previously, condition (4) is valid in the worst case situation, i_b being now the batch current.

3. RF drive generation

During the transient phase of the tuner, we must synthesize i_g' to meet condition (2) and correct for the effect of beam loading. It obviously implies that i_g' (or the corresponding power P') is available from the RF generator, otherwise transient beam loading cannot be corrected completely. To determine whether this is acceptable or not one must rely on simulation programs [6] [7]. Various techniques used to generate the proper i_g' will be examined in the following.

a) Amplitude and phase servo loops

The synthesis of i_g in such a way as to keep V constant irrespective of the beam loading can be done with two servo loops (Fig.5): the first, acting on the amplitude of i_g (amplitude loop) controls |V| and the second maintains the relative phase of V and i_b constant through the control of the phase of i_g (phase loop). The cut off frequency f_c of the loops must be much larger that the synchrotron frequency f_s, which means very strong damping of beam oscillations. This justifies the simplified stability analysis [4] [13] where the beam transfer function is neglected. f_c is obviously limited by the delays in the system, including the cavity bandwidth, but more fundamentally by f_o. The simple configuration of Fig. 5 with high loop gains cannot correct transient beam loading at f_o and its multiples.

Steady beam loading with its associated cavity detuning could excite mode n = 0 (Robinson instability [14]) if it were not heavily damped by the phase loop. However, mode n = 1 (one wavelength per turn), which is not damped may show up also and must be suppressed by a dedicated feedback circuitry acting through the RF cavity itself [15].

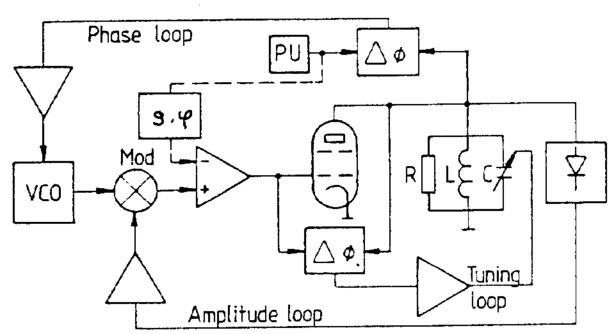

Fig. 5 Tuning, amplitude and phase loops
Feedforward correction (dotted line)

Independent amplitude and phase control of V is a well known technique for proton machines, it works satisfactorily for relatively small beam currents, i.e. when the gap voltage is predominantly determined by the generator current (typically $|i_b| < |i_g|$). For higher beam currents, a variation of the amplitude of i_g, for instance, not only results in a variation of the amplitude of V, but also of its phase. In other words, the two loops, which were independent at low beam currents become coupled together, and an unstable behaviour of the system results, above a certain beam current threshold. Pedersen's detailed analysis [4], confirmed by experiments on the CERN PS booster, lead to the generalized Robinson stability criterion, valid for Φ_s = 0 :

$$\frac{|i_b|}{|i_g|} < \sqrt{2 + \frac{f_a}{f_T} + \frac{f_T}{f_a} + \frac{f_p}{f_T} + \frac{f_T}{f_p} + \frac{f_a}{f_p} + \frac{f_p}{f_a}} \qquad (6)$$

where f_a, f_p and f_T are the unity gain frequencies of the loops (amplitude, phase and tuning respectively). Although the threshold is weakly dependent on the loop cut off frequencies, it might be dangerous in this configuration to increase the servo-tuner bandwidth.

Although it is, in principle possible to compensate loop coupling by an additional decoupling circuitry and increase the instability threshold [4], a much simpler solution is offered by feedforward correction.

b) Feedforward correction

With a pick-up electrode followed by a filter centered at f_{RF}, one can obtain a signal proportional to $-i_b$, independently from the RF system, and generate i_g' (RF drive with beam), according to (2) with a simple adder. Applied to the amplitude and phase servo loops described in 3 a, the method consists in injecting on the input of the RF amplifier the pick-up signal (with proper amplitude and phase $g.\varphi$) to generate the $-i_b$ current at the gap (Fig. 5). The amplitude and phase loops now act on the quantity i_g (corresponding to no beam loading) instead of i_g', and the cross couplings between loops are removed, as it can be shown analytically [16] and experimentally. As a result, the instability threshold can be considerably increased [16] [17] (stable operating conditions have been observed in the CERN PS for instance for $|i_b|/|i_g| \approx 8$ to 10).

The signal corresponding to $-i_b$ does not need to be synthesized with the ultimate precision, as it only removes the loop couplings and restores stability. For a varying RF frequency, the pick-up to cavity delay must be continously adjusted, and the variations in gain and phase of the RF power amplifier (assumed linear) corrected. In the CERN PS, a coarse feedforward correction (cavity compensation) covers the whole RF frequency swing during acceleration, but more precise settings are possible at a few critical (fixed frequency) points.

Feedforward correction can also be considered as a means to reduce the effective impedance of the cavity, seen by the beam. At the RF frequency, the beam induced voltage on the cavity amplifier combination is zero for a perfect correction. From this point of view, high amplitude and phase loop gains at f_s are no longer required to correct beam loading, as V is automatically kept constant by the feedforward compensation. Application of this technique (low loop gains) was for instance on the Brookhaven AGS, during adiabatic capture.

It is interesting to mention a variant of the feedforward technique, derived from the Alvarez linear accelerator technology. If the generator is a gridded tube (tetrode or triode), its output impedance is high if maximum RF power is to be extracted from the tube. When connected to the cavity by a long line, it fully reflects the beam loading wave travelling from the cavity to the generator. One can choose the length of the line to make the reflected wave cancel the beam induced voltage at the gap (the high impedance of the generator is then transformed into a quasi-short circuit, at the cavity) Note that, even with no voltage induced on the gap, the generator sees a mismatched load with beam, and must be able to deliver the current under this condition. This technique is in use on the CERN PS 200 MHz RF system, with trombones inserted on the feeder lines of the fixed tune cavities.

If the pick-up to cavity delay is adjusted to be exactly one turn (T_o), beam loading cancellation can be achieved, not only at f_{RF}, but also at frequencies $f_{RF} \pm n f_o$. This is relatively easy at fixed RF frequency, like in the CERN ISR [18], but with modern sampled or digital filters, variable delays, following a varying RF frequency can also be constructed [19]. The overall result is a rapidly changing impedance, ideally zero at frequencies $n f_o$, but twice as large at intermediate frequencies $(n+½)f_o$, where there are no beam current components. With a one turn delay and perfect cancellation, the voltage perturbation only lasts T_o which is small compared with T_s ($Q_s = (f_o T_s)^{-1}$ is usually « 1). In other words the reduction of the magnitude of the cavity impedance at the synchrotron satellites $n f_o \pm m f_s$ is also large (factor $(2\sin m\Pi Q_s)^{-1}$) for a small Q_s. The thresholds of the coupled bunch instabilities which can be induced by the cavity itself are not simply increased by that factor. As the phase of the residual impedance changes sign at $n f_o$, the complex synchrotron frequency shift curve is rotated in the dispersion diagram [19].

c) RF feedback around the power amplifier

We can consider the cavity itself as a beam pick-up tuned at f_{RF} and obtain the $-i_b$ signal from the gap. This leads to the configuration of Fig. 6, in which one obviously recognizes a feedback loop built around the RF power amplifier [20]. From the loop equations one obtains:

$$i_g' = i_g - \frac{GZ\, i_b}{1 + GZ} \qquad (7)$$

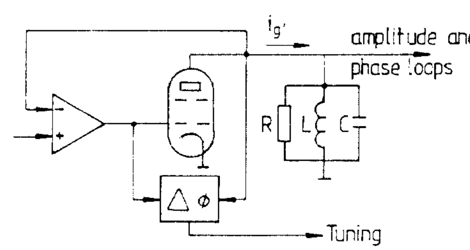

Fig. 6 RF feedback around the power amplifier

which, for GZ »1 (GZ: loop gain, Z cavity impedance) reduces to equation (2):

$$i_g' \approx i_g - i_b$$

i_g being here the generator current with no beam.

The feedback loop automatically generates the correct compensating signal, which is another way of saying that it keeps the controlled parameter V constant. One can consider RF feedback as a means to reduce the output impedance of the RF amplifier. A well known design is the cathode follower [21] with its low output impedance which shunts the cavity. Stability of the cathode follower with a reactive load needs careful study.

Even simpler, but of limited efficiency is to use a triode instead of a tetrode as RF power tube. The internal plate to grid feedback reduces the output impedance. In the same direction pulsing the DC current of the RF tube or powering second tube in parallel [22] has been used to reduce the output impedance of the RF amplifier for short periods.

In the case of Fig. 6, the cavity parameters (pole at $f_{RF}/2Q$) and the total delay of the feedback path determine the loop stability. The preamplifiers which are selected for the shortest propagation delay must be located very close to the power amplifier - cavity combination. As an example table I gives the parameters for the CERN PS booster second harmonic system, operating between 6 and 16 MHz [23, 25]:

Table I

Preamplifier gain	25 x
Bandwidth	150 MHz
Power	130 W (1 dB compression)
Propagation delay	5 ns
Impedance reduction factor:	21 dB at 6 MHz
	14,5 dB at 16MHz

For a varying RF frequency, one could, in principle adjust the delay of the return path to keep the 180° phase condition at f_{RF}. However, in many designs (2nd harmonic PS booster, future PS RF system, for instance) the total delay is kept short enough (wide bandwidth preamplifier)to ensure stability over the entire RF frequency range, without programming the phase. In this case it is extremely important to damp the higher resonances of the cavity (or to reject the corresponding signals) in order to avoid parasitic oscillations of the feedback system at high frequencies.

This technique is very attractive, as it reduces the effective impedance of the cavity not only at the RF frequency, but also over a large bandwidth. This feature is particularly helpful to avoid self bunching instabilities in storage rings for debunched beams (CERN ISR or AA for instance [20, 24]).

In such a conventional feedback system the total phase slip should be less than about $\pm \Pi/4$ over the unity gain bandwidth $2\ \Delta\omega$ of the system, giving the condition:

$$\Delta\omega = \Pi/4T \qquad (8)$$

where T is the overall delay in the feedback path. For a fixed tuned cavity and a small detuning angle, the cavity impedance, far from the ω_o resonance is approximated by $Z = R/2jQ(\Delta\omega/\omega_o)$. The overall loop gain GZ, at the $\pm\ \Delta\omega$ points is of the order of unity: this gives an upper limit for GZ and a minimum value of the impedance seen by the beam, R_{min} given by :

$$R_{min} = \frac{2}{\Pi} T \frac{R\omega_o}{Q} = \frac{2}{\Pi} \frac{T}{C} \qquad (9)$$

The ultimate performance of wideband RF feedback only depends on T and the cavity geometry.
If applicable (small T) this is the best solution to the problem of beam loading: wide band coverage and no need for critical adjustments.

Very large impedance reduction factors (several orders of magnitude) have been achieved at low RF voltages (CERN AA for instance) and fixed cavity tune (no servo loop). With a tuning servo the usual phase discriminator measuring the control grid to gap phase difference (Fig. 6) may cease to work properly if the active power delivered by the beam to the cavity exceeds the amplifier active power. This is the case for $\Phi_S \simeq 0$ and very high beam loading factors or if counterphasing of several cavities is used, which completely changes the phasor diagram of Fig. 2. A possible solution is given to this problem by measuring the normalized reactive power of the amplifier [25].

d) The RF feedback with long delay

In large RF systems (CERN SPS for instance), long delays may be unavoidable, and the conventional RF feedback would have a too restricted bandwidth (much smaller than the cavity bandwidth itself in the SPS case). Transient beam loading at multiples of f_o would not be corrected,leading to phase oscillations of fractions of the beam and possibly coupled bunch instabilities.

In order to solve the problem, we observe that a large gain G is only needed in the vicinity of the revolution frequency harmonics, where beam current components exist. Outside these bands, the phase rotation due to the excessive delay will be unimportant if G can be made small enough. With a return path transfer function having a comb filter shape with maxima at every f_o harmonic, this condition can be satisfied. In addition, the overall delay of the system must be extended to exactly one machine turn (T_o) to ensure a zero phase at the $f_{RF} + n\ f_o$ frequencies.

The comb filter transfer function is of the form:

$$H\ (j\omega) = \frac{G_o}{1 - K \exp\ (-j\delta\omega T_o)} \qquad (10)$$

where G_o and K are constants ($0<K<1$).

Combined with the one turn delay (transfer function: $\exp(-j\delta\omega T_o)$), the overall open loop transfer function becomes:

$$G\ (j\omega)Z(j\omega) = \frac{G_o Z(j\omega)}{\exp\ (j\delta\omega T_o) - K} \qquad (11)$$

represented in the complex plane by a circle, for a slowly varying $Z(j\omega)$. The complex plane origin is encircled and therefore the gain of the system is limited by the stability condition. In the vicinity of the cavity resonance, where Z is maximum and real (note that for a travelling wave structure Z is always real [5][16]) the circle crosses the negative real axis at a distance $-G_o/(1+K)$ from the origin. Stability requires obviously $|G_o Z|<1+K$; it can be shown that this condition is also sufficient even outside resonance for an RF cavity approximated by a single RLC equivalent circuit.

The apparent impedance of the cavity Z':

$$Z' = Z \frac{\exp\ (j\delta\omega T_o) - K}{\exp\ (j\delta\omega T_o) - K - G_o Z} \qquad (12)$$

is real for frequencies:

$$f_{RF} + n\ f_o \quad ; \quad Z' = Z \frac{1 - K}{1 - K + G_o Z} << Z \qquad (13)$$

and:

$$f_{RF} + (n+\tfrac{1}{2}) f_o \quad ; \quad Z' = Z \frac{1 + K}{1 + K - G_o Z} \qquad (14)$$

To stay at a reasonable distance of the stability limit, take for instance $G_o Z = (1+K)/2$.

This gives, at frequencies $f_{RF} + (n+½)\ f_o$ $Z' = 2Z$, as in the case of feedforward correction, whereas for the revolution frequency harmonics, one obtains:

$$Z' \simeq Z\ (1-K) \qquad (16)$$

for $(1-K) \ll 1$.

By making K close to unity, RF feedback approaches the theoretical performance of the feedforward correction, but with all the inherent advantages of closed loop systems (no critical adjustments). Similarly the time response of the RF feedback is entirely determined by the one turn delay, as in the feedforward case (note that the unity gain frequency of the servo is of the order of $f_0/2$). This is confirmed by the observed transient response of the SPS system.

The residual impedance at the synchrotron sidebands is approximately the same as with a one turn delay feedforward correction (for K=1 and $G_0Z=1$); its phase changes sign at each n f_0 harmonic, which results in a rotation of the complex synchrotron frequency shift curve. The coupled bunch, cavity driven, instability thresholds must be obtained numerically [19].

Except for relatively small machines with fixed RF frequency, long delay feedforward or feedback techniques could only be envisaged with the help of modern signal processing technology, i.e. sampled or digital filters. The digital comb filter is derived from the well known first order low pass recursive filter. With a sampling frequency Nf_0 (locked to a subharmonic of the RF frequency), the theoretical bandwidth of the filter is N $f_0/2$, corresponding to N/2 maxima in the comb filter response (N = 462 in the SPS design). Implementation of the one turn delay is straight forward in digital technology with a memory (R.A.M. or first-in. first out type).

The speed of the various elements, limited by the cycle time (T_0/N) may become very critical, requiring the fastest A-D converters (flash converters), memories and multipliers (parallel multipliers). For this reason the number of bits is limited: 8 bits in the ADC, 12 bits in the multiplier array, in the case of the SPS, but no adverse effects from the quantization errors can be observed. However K cannot be made very close to unity, with a small number of bits: the residual impedance Z' at the revolution harmonics is essentially determined by this technological limitation (1-K = 1/8 for the SPS case).

The RF signals may have to be translated in frequency to be conveniently processed. Coherent mixing with separate channels for in phase and quadrature components is necessary to reject the unwanted image frequencies (measured rejection >35dB), and make the overall electronic chain look a linear system. For a varying RF frequency the correct phase can even be maintained with an artificial delay inserted between the output and input local oscillators (Fig. 7).

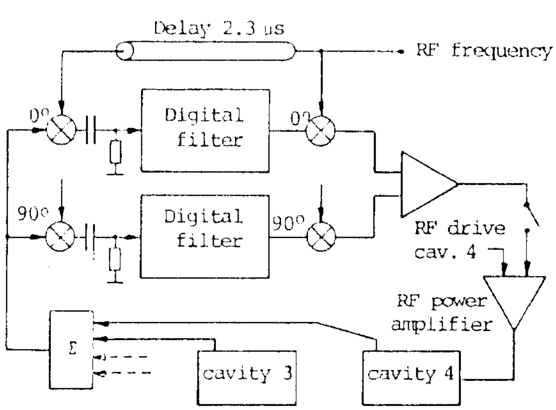

Fig. 7 Layout of the RF feedback system

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