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THE PS 10 MHZ CAVITY AND POWER AMPLIFIER

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1 PREAMBLE

The PS 10 MHz cavities in straight sections 11, 36,46,51,56,66,76,81,86,91 and 96 (10 cavities plus the spare in SS11) are vital to the functioning of the PS. It is these cavities which form the bunches after injection, accelerate them to the desired energy and perform gymnastics such as bunch splitting and rotation prior to ejection to other rings or users. The cavities are programmable both in gap voltage (0.5 kVp to 10 kVp), in frequency (2.8 MHz to 10.01 MHz) and in the way they work in groups during the PS super-cycle. Each programmed cavity group is tuned correctly to the low level RF signal by partial saturation of the cavity ferrites via an automatically adjusted current.

A cavity consists of two ferrite loaded ¼ wave lines with the capacitive gap (730 pF equivalent) at the input end. The gaps are in series for the beam but are fed in parallel from the RF power amplifier. The gap loss resistance is set mainly by the ferrite losses and the power amplifier output impedance. The cavity ferrite loss resistance alone is quite high (approx 22k ohms/gap) for low level signals at LF (3 MHz) giving a cavity unloaded Q of 130 which decreases to 30 when the final amplifier is connected, due to the tube internal resistance of about 3.5 kΩ (Ia = 1.5 A). At high gap voltage the ferrite losses increase further, reducing the loss resistance to about 5kΩ/gap at 10 kV, 10 MHz (loss measurement by cooling water temperature). The total cavity impedance seen by the beam under these conditions is therefore about 6kΩ.

As seen from each cavity gap a bunched beam in the PS ring constitutes a constant current source with a frequency spectrum which depends on the bunch width, the spacing between bunches and the number of particles in each bunch. Basically the DC component of this current Idc can be calculated from the total charge (number of particles) passing the gap in the bunch circulating time Tr. With intense bunches of width Tw the current harmonics at intervals 1/Tr can have high amplitudes up to a frequency 1/Tw and will induce large transient voltages in the gaps, which may disturb the circulating beams. At low power amplifier drive levels the induced beam voltage can be higher than that produced by the amplifier and can cause system instability.

To reduce the beam-induced voltages either an anti-phase signal can be derived from the beam and fed into the power amplifier (known as feed-forward) or, better, the anti-phase cavity gap voltage itself (fast feedback (FB)). The latter method is the classical one where the output impedance of the power amplifier (3.5 kΩ and less at higher drive levels) is reduced by the loop gain of the feedback loop. All the other advantages and disadvantages of feedback such as reduced distortion and drift, possible loop instability and the need for higher input drive power are also present.

The RF harmonics from the beam, in particular the harmonic at the frequency to which the cavity is tuned, are not in phase with the RF drive program and the amplifier feedback, in working to reduce what it sees as distortion, causes the power level in the amplifier to rise. This is generally known as Beam Loading. The feedback loop will be stable providing the loop gain remains below 0dB, at frequencies above and below f_0, where the loop phase shift reaches 180 deg (in addition, of course, to the 180 deg necessary to make the FB negative). The resonant cavity alone produces 90 deg, leaving 90 deg for the phase shift in the amplifier (for good response in fact only about 40 deg is tolerable). The Q of circuits in the amplifier and the cable delays must therefore be kept as low as possible.

At present the feedback applied to the 10 MHz amplifiers has a loop gain of the order of 20dB, i.e. a factor 10 leading to an amplifier output impedance of the order of 350 Ω/gap (measured 320 ohms). At high beam intensities the induced gap voltages can still cause some degradation of the circulating beams. A further reduction in cavity impedance has been achieved by the addition in 1995 of the '1 turn feedback' system [1]. This feedback loop achieves an effective further 12dB reduction of cavity-induced voltage by acting only the harmonics spaced 1/Tr. A comb filter in the loop is used to reduce the gain to zero at all frequencies which could cause instability i.e. those not at the harmonic frequencies where the loop phase is 180 deg.

Under certain conditions during beam gymnastics there is zero drive to the power amplifier and only a minimum of induced gap voltage is acceptable. In this case programmed high voltage vacuum relays with low resistance damping networks are used to reduce gap impedance to a few ohms. The relay contacts are often required to open or close with the full bunched beam current flowing, a duty known as 'hot switching'. The relay lifetime under these conditions is short and is a major cause of system failures in the 10 MHz RF system.

2 DETAILED DESCRIPTION OF THE STANDARD PS 10MHZ AMPLIFIER

Refer to the schematic diagram of the amplifier PS/RF-HC 3044/7. Three separate stages of amplification can be identified. The first YL1056, T1 whose control grid (g1) forms the summing point for the feedback (FB) signal and the RF input from the Herfurth surface driver amplifier is known as the 'Pre-Driver'. The second two YL1056 tubes T2 and T3 in parallel, the 'Driver', are coupled to the ' Final' output tube T4 grid via a tunable resonator TR1. Local, or internal feedback is applied from the TR1 load resistor to the summing junction so as to reduce the 'Q' of TR1, whilst overall feedback is taken from a capacitor disc pickup on the Final anode (C52). A simplified block diagram of the amplifier would appear therefore as in Fig. 1.

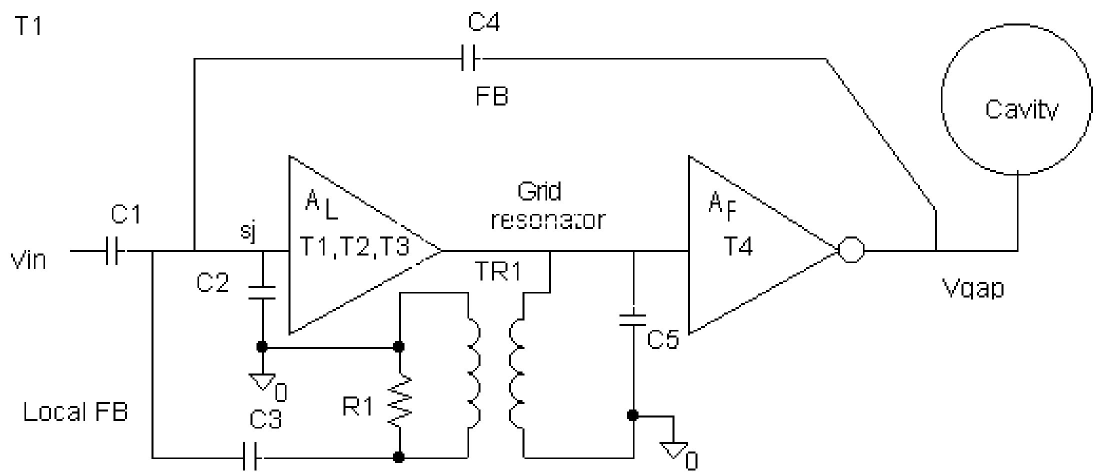


Fig. 1a : Simplified Amplifier circuit

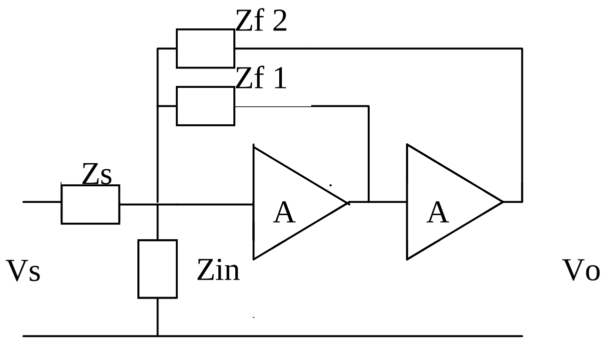


Fig. 1b : A further simplification

Referring to Fig. 1b some expressions for the main amplifier parameters can be stated.

The loop gain of the local feedback, $AB_1 = \dfrac{A_1 \cdot ZT}{Zf_1 + ZT}$ where $ZT = Zs \,//\, Zin$

The Loop Gain (LG) of the overall feedback, $AB_2 = \dfrac{A_1 \cdot A_2 \cdot ZT_2}{Zf_2 + ZT_2}$ where $ZT_2 = \dfrac{ZT}{1 + AB_1}$

The overall Closed Loop Gain (CLG) $= \dfrac{Vo}{Vs} = \dfrac{Zf_2}{Zs} \cdot \dfrac{AB_2}{1 + AB_2}$

The Open Loop Gain (OLG) $= \dfrac{Vo}{Vs}(Zf_2 = \infty) = \dfrac{Zin}{Zs + Zin} \cdot A_1 \cdot A_2$

The effect of the local loop, apart from decreasing the output Z of the local amplifier A1, is to reduce the summing junction impedance ZT such that the LG decreases. Providing the LG>>1 one can also state that LG =OLG/CLG.

2.1 Pre-Driver stage

Here a single Siemens YL 1056 air-cooled tetrode operates in class A. The typical transfer function from g1 Pre-Driver to g1 Driver, i.e. the gain and phase of the first stage, is shown in Fig. 2a and Fig. 2b.

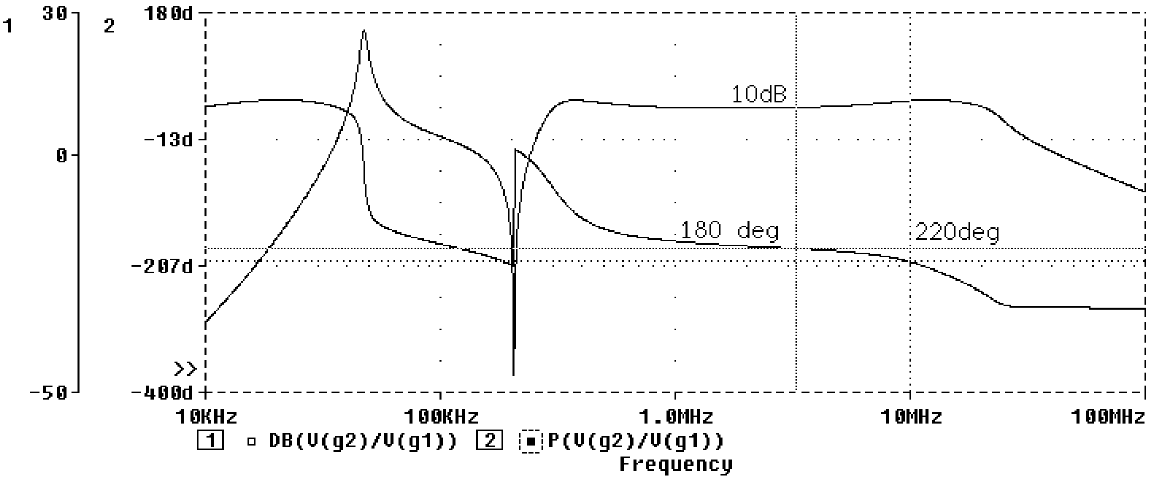


Fig. 2a: Gain and Phase of Pre-Driver stage (Spice simulation)

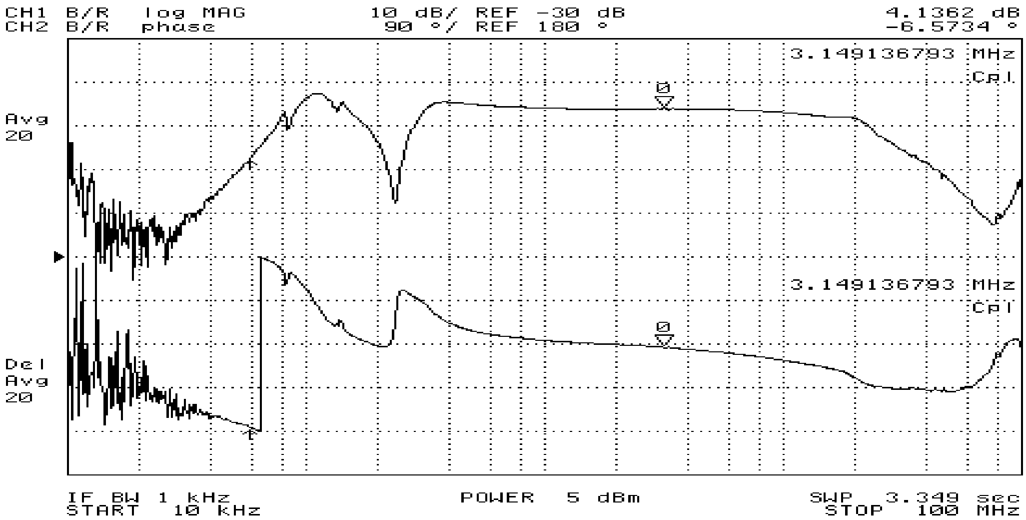


Fig. 2b : Gain and Phase of Pre-Driver stage (NWA Measurement)

The plot Fig. 2a was obtained from the Pspice simulation program (10 MHzamp) where the program Schematic has been adapted so as to reproduce very closely the plot Fig. 2b, obtained from measurements on the amplifier using a Network Analyser. Between 0.4 MHz and 20 MHz the gain is essentially flat. The phase shift of this stage between 3 and 10 MHz (>-20 deg.) sets an upper limit on the stability of the amplifier FB loop. For a given tuning range it determines the minimum Q (and therefore the group delay) needed in the Driver stage resonator which is detuned above fo at 10 MHz so as to bring the loop phase back to 180 deg. The anode load of the Pre-Driver consists of the Cg1k of the Driver stage in parallel with the 47 µH anode choke and the 50 ohm load resistor R in series with the compensating 0.52 µH inductor L. This inductor resonates with the stray load capacities at about 12 MHz and thus improves the response at 10 MHz and, more important, reduces the stage phase shift between 9 MHz and 11 MHz by a few degrees. A resonance at 270 kHz due to the Pre-Driver anode choke and the supply filter is visible.

As seen from the plots, the stage gain with 800mA Ia bias current is about 10dB which corresponds to the tube gm of 60mA/Vx 50 ohms.

2.2 The Driver stage

Two tetrodes are connected in parallel for added gain and power output. The transfer function of this stage depends on the resonant frequency of the final g1 coupling resonator. The gain and phase from g1 Driver (g2) to g1 Final (gf) at a fo of 3 MHz is shown in Fig. 3.

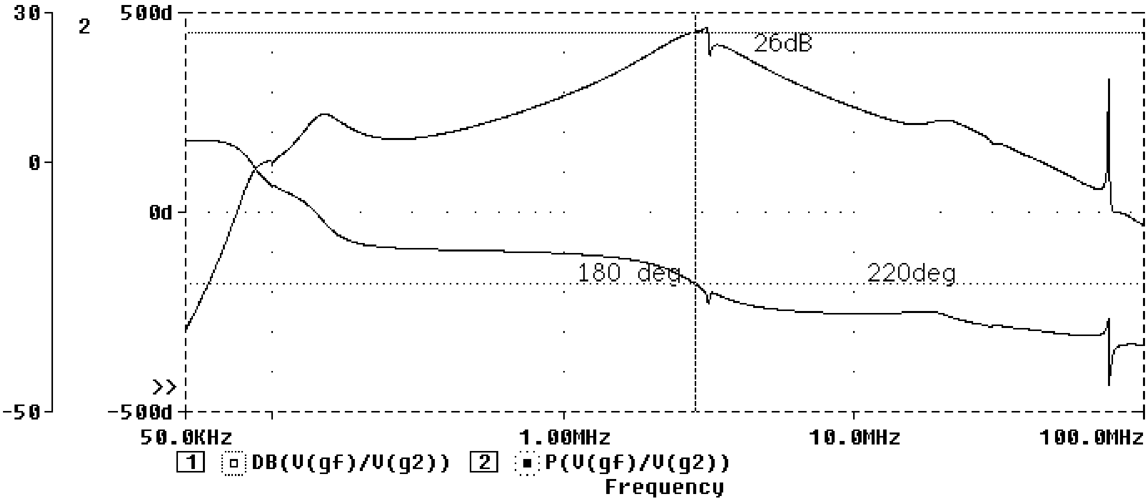


Fig. 3a : Gain and Phase of Driver stage at fo = 3 MHz (Pspice simulation)

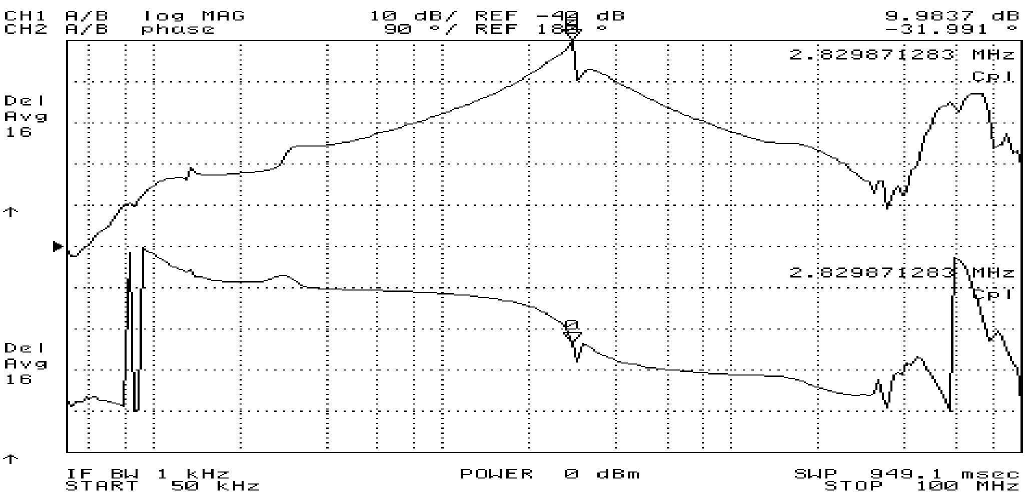


Fig. 3b : Gain and Phase of Driver stage (NWA measurement) fo = 3 MHz

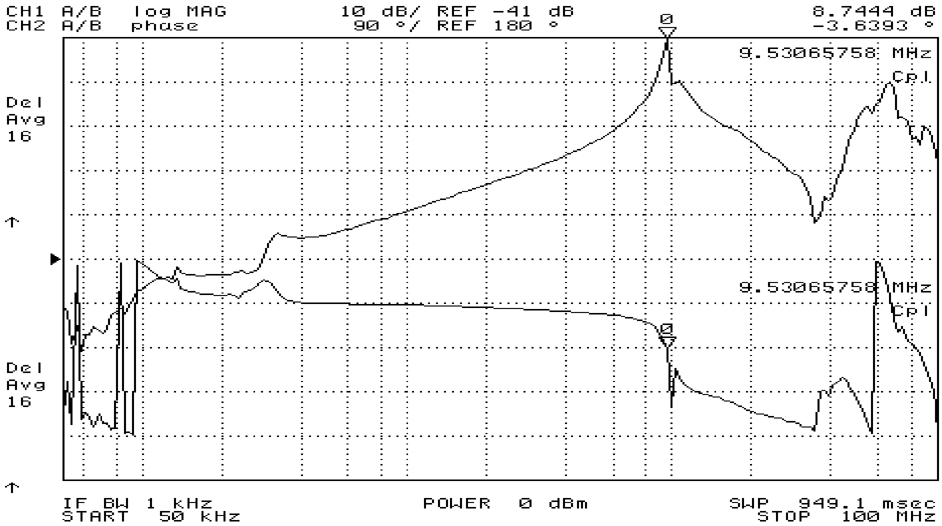


Fig. 3c : Gain and Phase of Driver stage (NWA measurement) fo = 10 MHz

The plot in Fig. 3a was obtained from the Pspice program 10 MHzamp2.5 m where the cavity fo is 3 MHz.

Miller capacity in T4 causes feedback to the grid around the cavity fo as can be seen particularly in Fig 3c. It can be deduced from the simulation program that the Cag1 capacity (which is at the root of the Miller effect) is effectively slightly higher than the 1.9pF given in the tube data book. This is probably due to imperfect decoupling of T4 g2 (screen grid) to RF ground and to the g2/g1 coupling on the socket. The effect of the Miller capacity is mainly to make the Final stage gain more sensitive to the grid resonator tuning (see Annex 1).

The local FB capacitor (C39, ~ 6pF) produces a local loop gain of about -2dB, reducing the closed loop Q of the final grid resonator from 2.5 to 1.3 and from 6 to 3.8 at 3 MHz and 10 MHz respectively. At an Ia bias of 400mA/tube the stage operates in class AB. The stage gain at fo = 3 MHz is seen to be about 26 dB which corresponds to100 mA/V T2+T3 gm and the 200 ohms gf load resulting from the 50 ohm external load R33 stepped up via the 2:1 ratio resonant transformer TR1.

This autotransformer is of special construction. The primary is formed by two copper sheet ‘half doughnuts’, split and connected in ‘figure of 8’ style so as to form a two turn winding on a core composed of two 4L2 ferrite rings. The flux due to this winding flows in opposite directions in each ferrite ring and therefore does not couple to the 50 turn ferrite saturation winding which is wound over the whole assembly

The advantage of the ‘half doughnut’ construction is that the ferrite RF flux is constrained inside the copper sheet, thus reducing the leakage flux and, consequently, resonator inductance at higher frequencies when the ferrite is operating at near saturation. TR2 is a ‘balun’ which isolates the 1-turn secondary picked off the centre of the primary, thus allowing the connection to the 50 ohm load to be reversed so that the local feedback via C39 is negative. In spite of the low initial ferrite ‘mu’ of 170 and the high stray capacity of the sheet primary winding a max to min inductance ratio of about 10 and a maximum fo of 10 MHz is achievable with 20A in the tuning winding.

2.3 Final stage T4

This stage develops the power needed to drive the cavity and beam loading losses. A Siemens RS1084 CJ water-cooled tetrode operates in class AB at an anode bias current of 1.5A and a DC anode supply voltage of 15 kV. Demineralized water is supplied to the anode through 40cm of high pressure insulating pipe at 30 lit/min. to keep the anode outlet water temperature to below 50 deg C at maximum anode dissipation. The typical transfer characteristic of this stage from T4 grid to anode and for a cavity fo of 3 MHz is shown in Fig. 4 below.

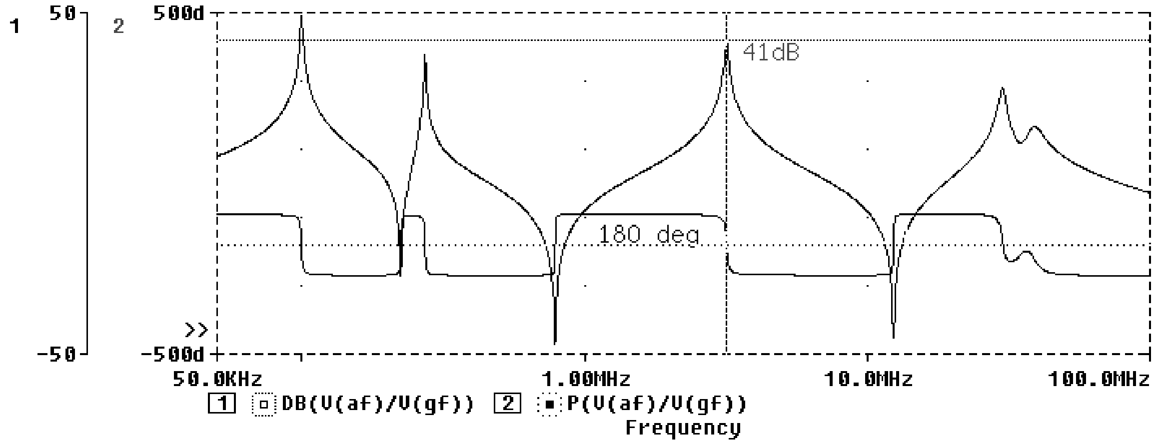


Fig. 4a : Gain and Phase of Final stage at fo = 3 MHz (from Pspice simulation)

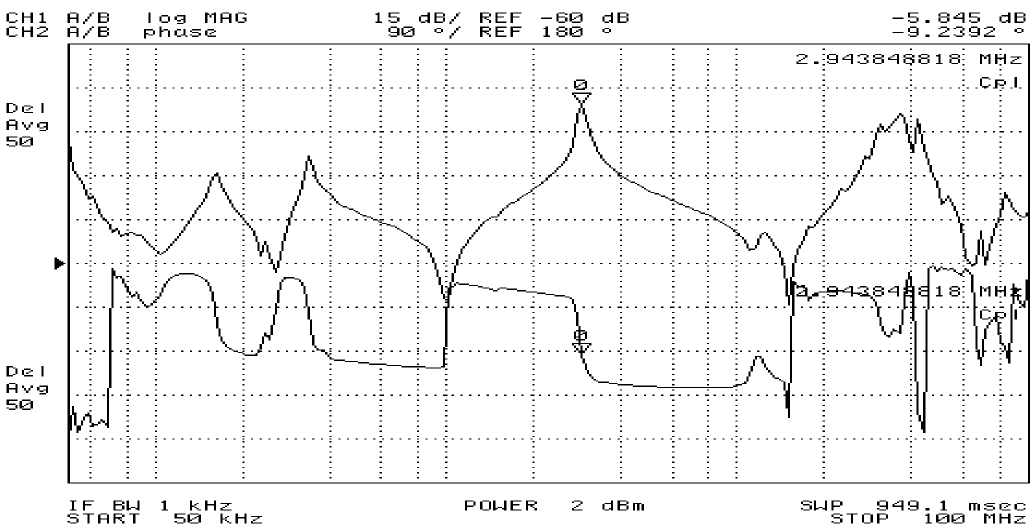


Fig. 4b : Gain and Phase of Final stage at fo = 3 MHz (NWA measurement)

The plot of Fig 4b is of some relevance since it is from T4 anode that the overall feedback is derived and not from the cavity gap. The gain peaks at about 40dB. It will be seen that the phase at frequencies either side of fo reverses at the zeroes around 1 MHz and 14 MHz, which are due respectively to the 10nF coupling capacitor and the 260 nH inductance of the coupling line from the tube anode to the cavity gaps. At 280 kHz the resonance due to L17, the anode choke, with C57 and C58 is important since the Pre-driver gain is still high at this frequency. Above 30 MHz the anode capacity and the coupling line resonate strongly and are damped by the 35 MHz anode band pass filter to R59.

3 THE LOOP GAIN AND AMPLIFIER STABILITY

A relative loop gain and phase measurement is made by dividing the Final anode output pickup signal by a signal injected into the summing point at T1, g1 via a 20pF capacitor. The contribution at the summing junction of the local FB fromTR1 in addition to the main FB from T4 anode is then accounted for.

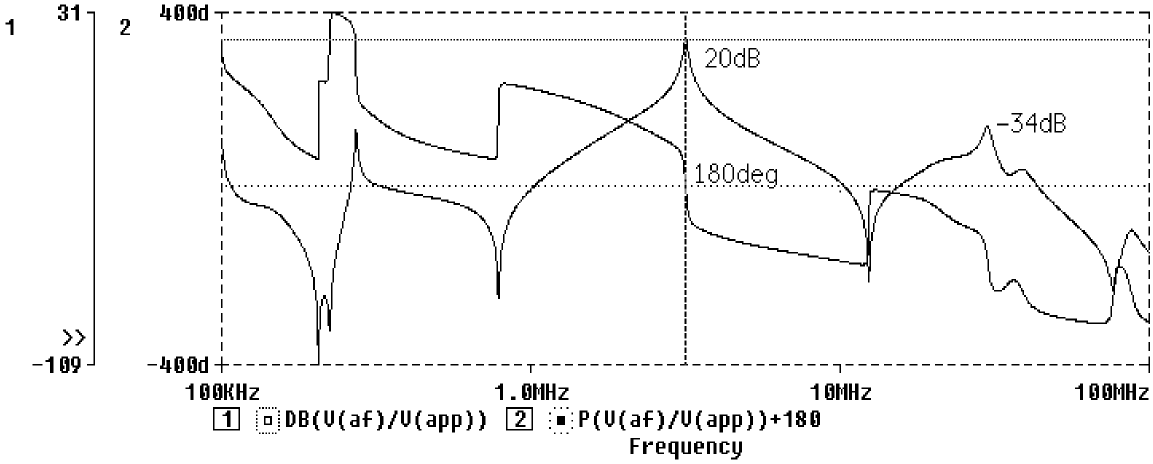


Fig. 5a : Typical loop gain and phase at fo = 3 MHz (Pspice simulation)

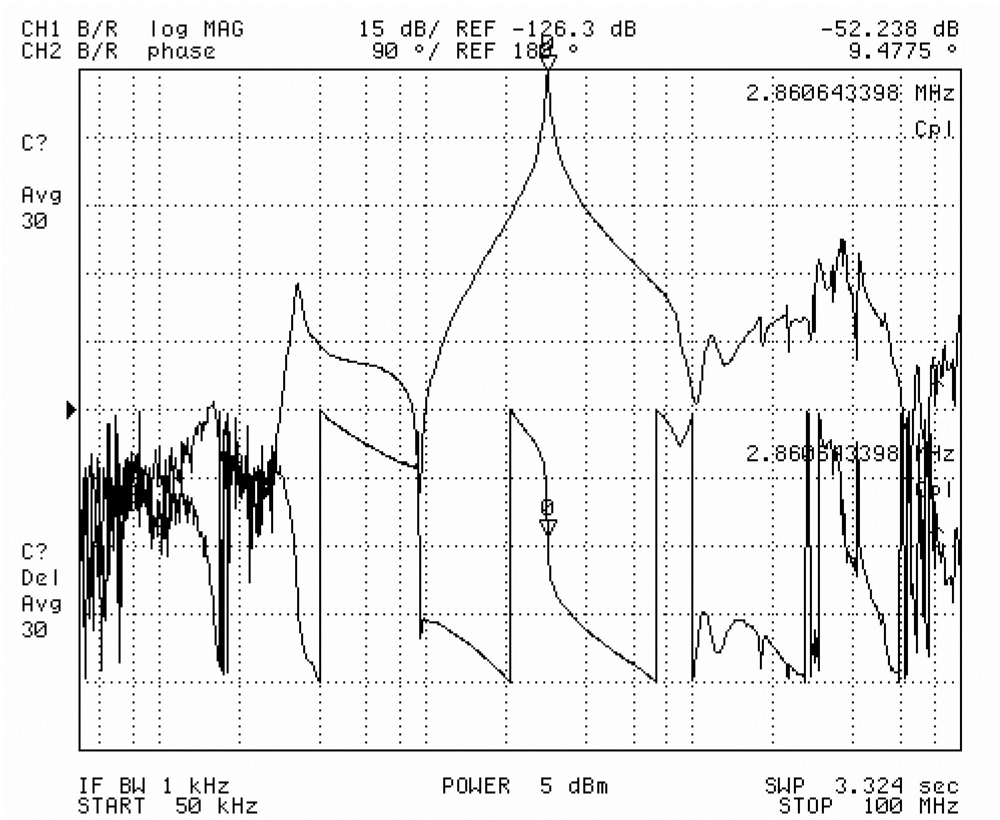


Fig. 5b : Typical loop gain and phase at fo = 3 MHz (NWA measurement)

The typical loop gain and phase at 3 MHz fo is shown in Figs. 5a and 5b. The loop characteristic is obtained from Vaf/Vapp. where Vapp is the signal applied to the 20pF dummy final anode feedback capacitor. This method of overall loop gain measurement is necessary since the g1 summing junction impedance is modified by the loop gain of the local FB and cannot be simulated by a passive dummy network. At ~3 MHz the closed loop response can be seen to be stable around f_0 since the phase of the loop including FB from Final anode to T1, g1 has barely turned by 90 deg at the points where the loop gain is zero dB. At 10 MHz fo the phase rotation of the total FB at the zero dB points is higher than 100 deg. due to the group delay of the grid resonator (Q with local FB=3.8, group delay at 10 MHz = 90 ns) and the limited bandwidth of the Pre-Driver stage. The closed loop is stable up to 26dB of loop gain at 10 MHz but there is strong gain distortion at 2.8 MHz and 11 MHz due to the loop phase shift.

There is some doubt about the relatively high loop gain seen at 32MHz. The final anode signal amplitude is very large at this frequency and some noise breakthrough on the cables of the highly attenuated anode pickup is possible. Secondly a low gain margin would result in an increase of output signal when the loop is closed and this does not happen to any extent.

C42 transforms R33 via the balun cable to give a small phase advance and a slightly lower Driver stage gain at 10 MHz, which improves the gain margin at this frequency in spite of a lower local LG and therefore a higher grid resonator Q. At 285 kHz the T4 anode resonance is visible but well below the zero dB line, since R5 on the Pre-Driver grid causes strong loop attenuation at low frequencies. However, R5 creates phase advance at LF and a loop gain limit arises when the cavity fo is at minimum (2.6 MHz). The phase margin is then small so that the overall gain starts to rise (the 'surtension' at 20dB LG mentioned in the test procedure). If the loop gain is raised above 27dB instability occurs at this frequency. At 30 MHz the anode to cavity line resonance at the Final anode is evident but is strongly attenuated by the low cutoff frequency of the Pre-driver and Driver responses.

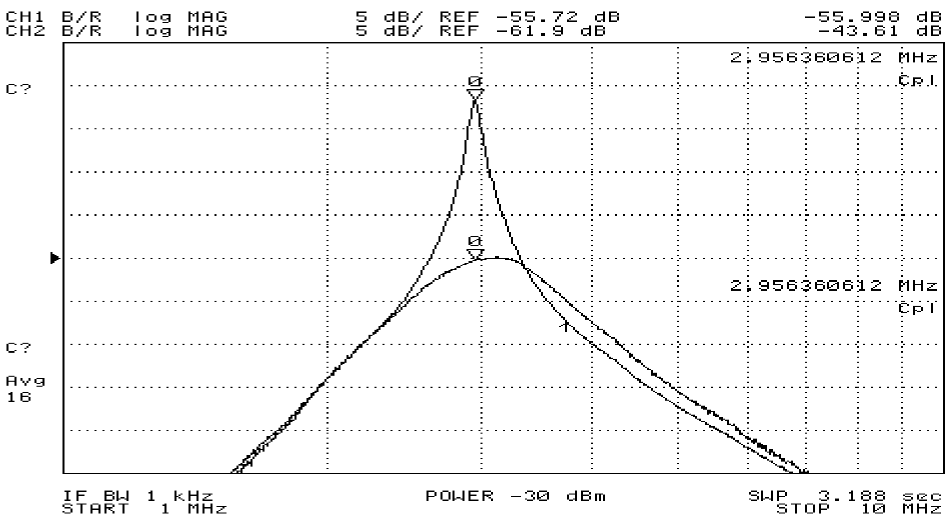


Fig. 6a : Open and closed loop response at 3 MHz (NWA measurement)

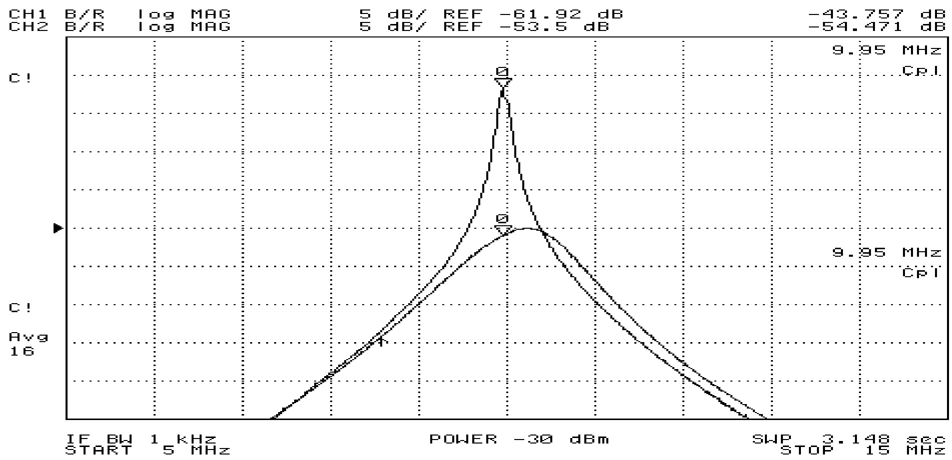


Fig. 6b: Open and closed loop response at 10 MHz (NWA measurement)

Figs. 6a and 6b show the reduction in cavity Q obtained with FB. In 6a the open loop cavity response with a Q of 38 is reduced to 3.53 when the loop is closed corresponding to a BW of 850 kHz. In Fig. 6b, the corresponding figures are 89 reduced to 11, corresponding to a BW of 900 kHz.

Off resonance, the closed loop gain is not symmetrical about the cavity fo. The cause is complex, depending not only on whether the final grid resonator constitutes capacitive or inductive coupling at a particular frequency but also on the deviation from 180 deg. of the loop phase at fo and the grid resonator fo offset. These mechanisms are further complicated by the effect of the FB and the loop group delay. Adjustment of the grid resonator-tuning program is necessary to minimize asymmetry.

To produce 10kVp at 3 MHz at the gap with an overall gain of 47dB requires 42Vp at the amplifier input, or 18 W into 50 ohms. This rises to about 23 W at 10 MHz.

4 MODIFICATION OF THE AMPLIFIER TO INCREASE LOOP GAIN

With the prospect of higher intensity beams in the PS there is some interest in obtaining a higher amplifier loop gain so as to decrease the cavity impedance to the beam. An increase of 6 to 10 dB in loop gain would be useful. Considering the critical importance of these cavities to the functioning of the PS and the large number involved, in the short term, it would be prudent to make only limited modifications.

A first obstacle to increasing loop gain is the reduced stability at 2.6 MHz when the nominal loop gain is increased by 6 dB. The loop still has several dB gain at this frequency when the cavity tuning currents are zero. The phase advance is >180 deg due to the 470 ohm (R5) on T1, g1 and the grid resonator which cannot be tuned below 2.8 MHz with zero ferrite saturation current. If R5 is increased to 1.5 k the loop phase advance is reduced and the phase margin becomes sufficient at 2.6 MHz. A second instability at 280kHz then appears where the resonance due to the Final anode HT supply filter is no longer attenuated by R5. This resonance can be damped at source by a damping network on C58 (see schematic). Increase of the Final anode FB plate capacitor to increase loop gain by 6dB is then possible providing the RF input power can be increased by a factor 4. Under these conditions with the cavity fo at

10 MHz, the ‘surtension’ at about 12 MHz due to phase distortion is 4dB, which is undesirable.

To reduce the phase shift above 10 MHz the Q of the grid resonator must be reduced. Measurements on this component reveal several stray resonances around 30-40 MHz which prevent increasing the local FB (see Fig. 3c). In addition, decreasing the local CLG would require a further increase in the RF input power for 10kV at the gap. A second problem is that even with the maximum current of 20A in the tuning winding the grid circuit does not resonate above 10 MHz on most amplifiers.

Reducing the inductance of the connection between the resonator winding and the Final tube grid made some improvement, since this modification simultaneously reduces the resonator Q and increases the resonant frequency. By this means the Q was reduced from 6 to 5 at 10 MHz, the fo raised from 9.9 MHz to 10.2 MHz and the peaking at 12 MHz from 4dB to 1dB. Further reduction of the phase lag requires an increase of the pre-driver bandwidth, but the resulting decrease of the gain would lower the local CLG which is unacceptable.

With the above modifications the loop gain has been increased by the expedient of increasing the feedback plate capacity which, though convenient, has the disadvantage of decreasing the closed loop gain. A loop gain increase of 6 dB then means four times the input power for the same gap voltage, or 95 W at 10 MHz. This is the maximum power at present obtainable from the surface ‘Herfurth’ amplifiers and at this level the 3^{rd} harmonic is only –10 dBC. At 10 MHz the 3^{rd} harmonic is 30 MHz, which falls close to the 32 MHz final anode resonance. So far during operation no damage has occurred to the damper resistor, so these modifications could constitute an acceptable compromise between better phase margins and a limited level of modification. The open and closed loop responses are shown in fig 7 below.

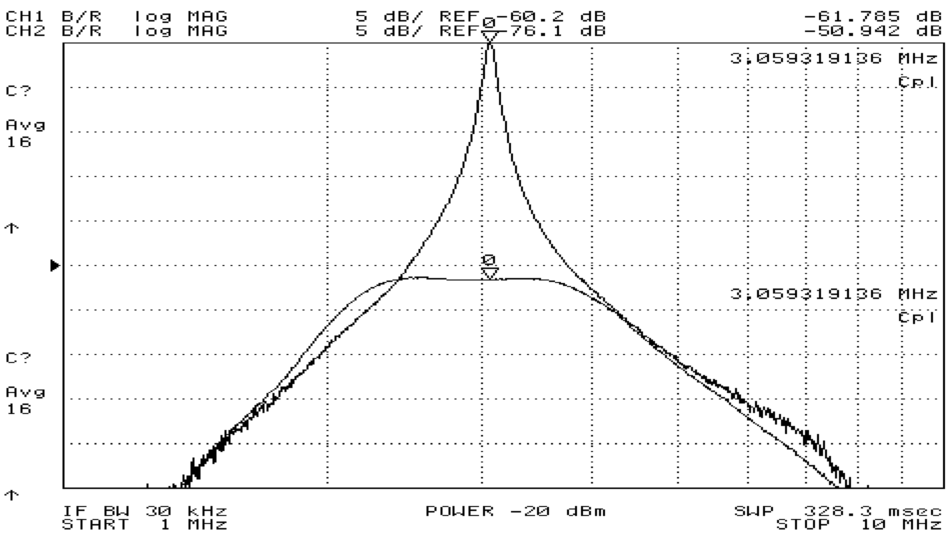


Fig. 7a : OLG and CLG at 3 MHz, modified amplifier, LG=26dB (NWA measurement)

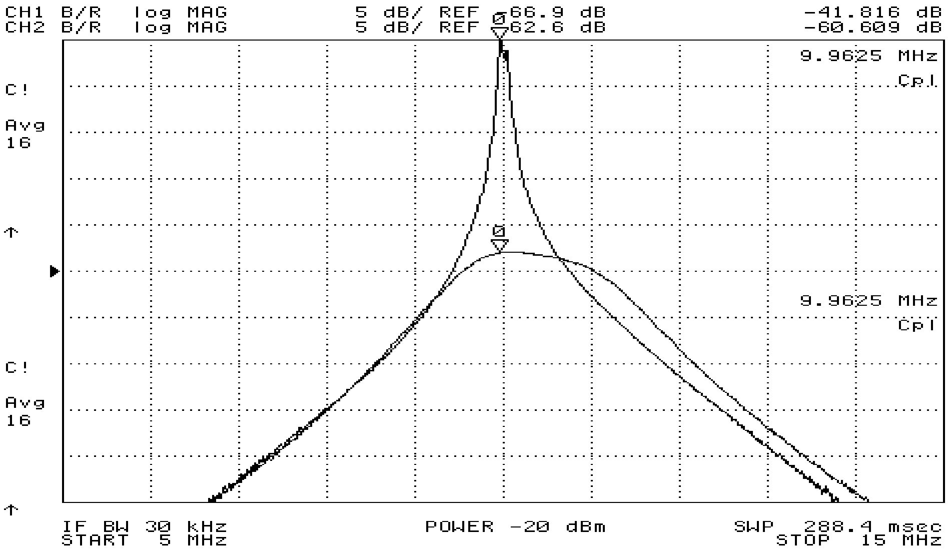


Fig. 7b : OLG and CLG at 10 MHz, modified amplifier, LG=26dB
(NWA measurement)

An important parameter is the effective cavity impedance seen by the beam with the amplifier feedback in operation. This is difficult to measure directly but the low level impedance seen on one gap, which is dominated by the amplifier output impedance, can be measured by injecting an RF current into one gap via a 3.3 k resistor and using a parallel resistor of 315 ohms or 150 ohms as a reference are shown in Fig. 8 below.

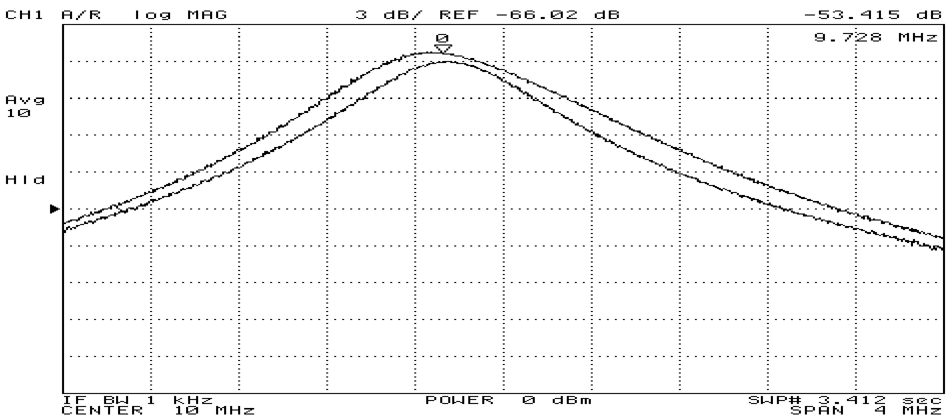


Fig. 8a : Gap impedance at fo=10 MHz with standard amplifier (LG=20dB). Lower trace is with 315 ohm reference resistor across the gap. NWA measurement

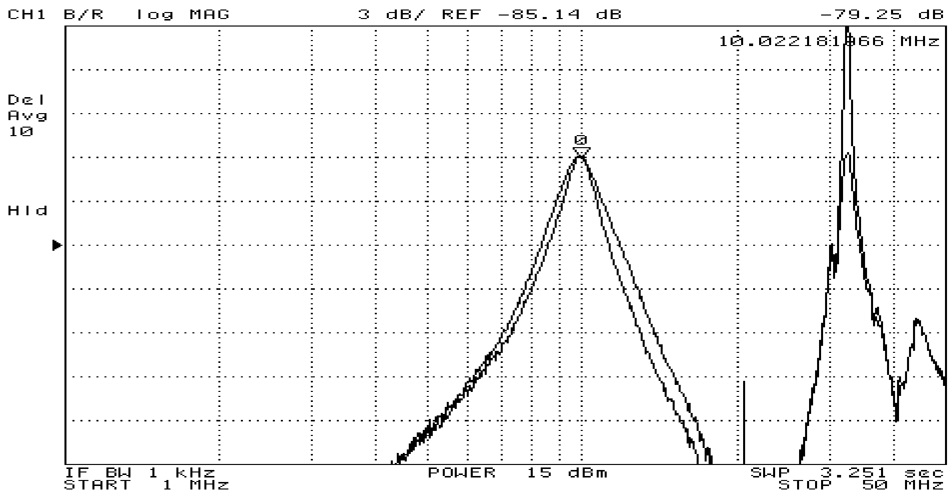


Fig. 8b : Gap impedance at fo=10 MHz with modified amplifier (LG=26dB). Lower trace is with 150 ohm reference resistor across the gap. NWA measurement.

The measurements show that the impedance seen at the gap is reduced by a factor 2 when the FB is increased by 6dB. It is about 320ohms with the standard amplifier and 150ohms with the modified amplifier. The figures for 3 MHz fo are similar. At off resonance frequencies the gap impedance is lower, but is naturally higher at the anode resonance around 32 MHz where 440ohms is seen with the standard amplifier (not shown) and about 580ohms with the modified amplifier.

As a further test the same measurement was made with the Final tube bias pulsed for 100ms from 1.5 to 4A. The gap impedance was 110 ohms during the pulse showing that the amplifier is stable with an added 4dB of loop gain. The cavity impedance seen by the beam would be roughly 4 times this gap impedance as measured by the above method.

5 INCREASE OF THE LOOP GAIN TO 30DB

In order to raise the loop gain to 30dB whilst maintaining the gain margins of the standard 20dB amplifier and a reasonable input power level (say 80Wmax) more extensive modifications would be necessary. The loop group delay would need to be decreased by a factor 3 and the OLG increased by 4dB (since we accept a decrease of the CLG by 6dB). Assuming the group delay can be decreased by a combination of decreasing the grid resonator stray capacity and connection inductance, increasing the local loop gain by 6dB and decreasing the phase shift in the Pre-driver around 10 MHz, then the Local forward gain (of the Pre-driver + Driver) must increase by 6+4=10dB and this without increasing the stray capacity on the summing point. If the stray capacity can be reduced here, ZT will increase and the Local gain increase can be scaled back proportionally.

One obvious weakness of the standard amplifier design is in the Pre-driver stage where the tube works into a 50 ohm load. This anode load is too low for reasonable gain and too high for low phase shift at 10 MHz. At maximum drive the anode swing is about 25 Vp whilst the anode supply is 800 V.

Tests were made using an anode-matching transformer but this was not successful (see Annex 2). The tube could be replaced by a Mosfet transistor[2] with higher transconductance, which would offer the increased gain x BW product required.

In order to test the feasibility of implementing such modifications the Pre-driver of one standard amplifier was modified to use an MRF 171 RF power Mosfet in place of the tube. The Mosfet has a transconductance of 1A/V, which is 20dB higher than the tube. Part of this gain can be exchanged for bandwidth by decreasing the output load.

After tests and measurements it was found that a reasonable compromise could be obtained by reducing the load from 50 to 15 ohms, giving 10dB more gain, and adding a 2nd input Mosfet (MRF 136) to buffer the high gate capacity of the gain stage thus reducing the gate capacity from 400pF to about 80pF(less than the original 120pF). This, along with a rebuild of the grid resonator, a separate Local FB inverter balun and the addition of damping networks to permit up to 10dB of local FB, has resulted in a 30dB loop gain and less than 70W input power for 10kV, 10 MHz at the gap. Fig 9a shows the CLG of the Local stages with the standard resonator at 5.5 MHz with –2dB of FB. A resonance at about 37 MHz limits the FB which can be applied.

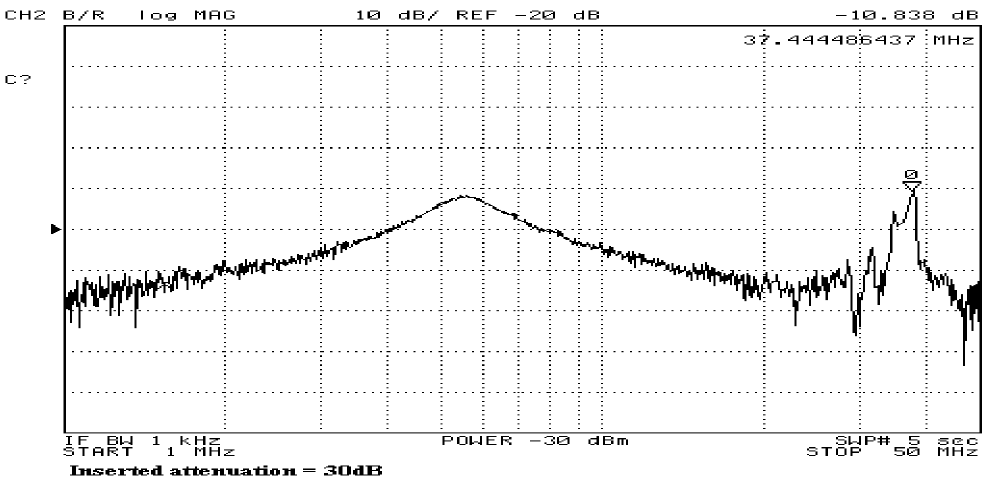


Fig. 9a : CLG of the standard Pre-driver + Driver with local FB.

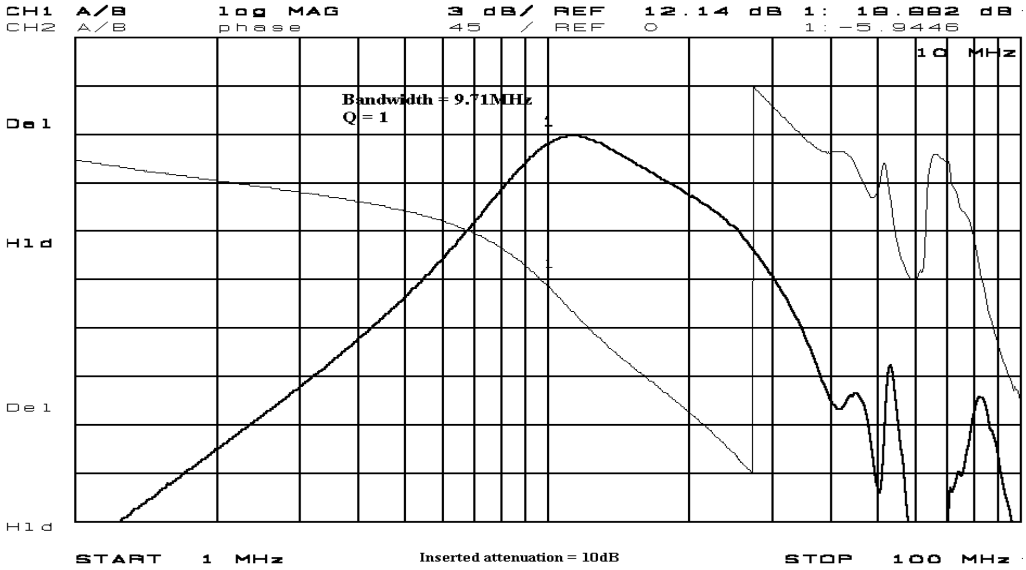


Fig. 9b : CLG of the Pre-driver + Driver with modified resonator and 10dB local FB.

In Fig. 9b the modified resonator is tuned to 10.4 MHz by 12A saturation current. The Q has been reduced from 4.7 to 1.1 by 10dB of Local loop gain. This was later reduced to 6dB to increase gain margin giving a Q of 1.6.

With 30dB overall loop gain the gain margin at the 32 MHz anode resonance was about 6dB which is only just sufficient. An improvement results if the FB is taken from the cavity gap via a 5pF HV capacitor feeding a short 3ns cable connected to the summing junction through a second 5pF. Fig 10 below shows the overall LG with FB taken from the anode and from the gap. An improvement of about 15dB at 32 MHz can be seen.

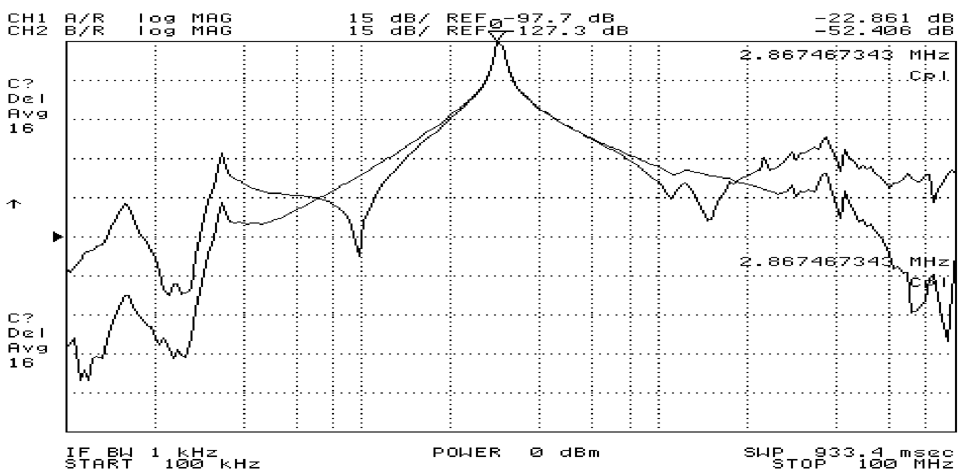


Fig. 10 : LG with FB from anode and from cavity gap.

An inverting balun fed from a Final anode pickup and with the inverted end connected to the final grid via a small capacitor was successful at neutralizing the Miller effect. At above 50 MHz however some resonances caused difficulties at high power operation and this circuit was finally removed.

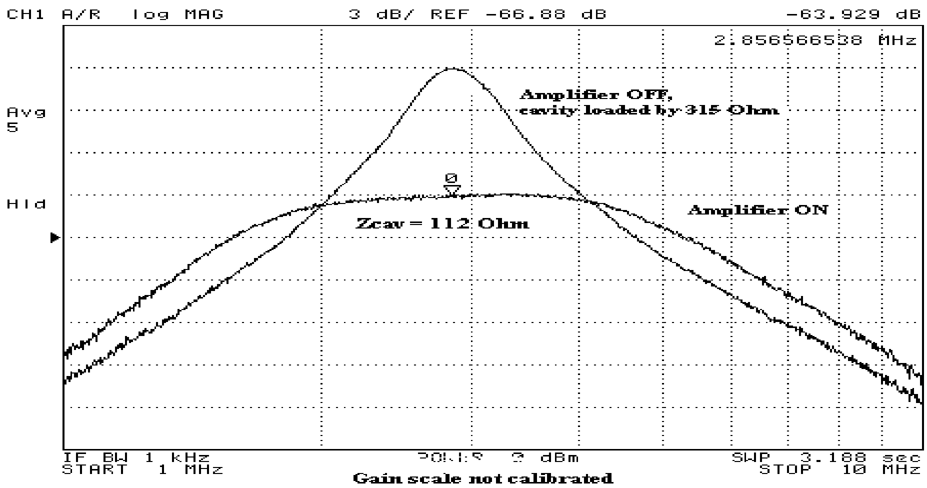


Fig. 11 : Cavity gap impedance at 2.85 MHz with 30dB loop gain. The impedance is reduced from 320 ohms with 20dB LG to 112 ohms with 30dB LG.

An increase in the dissipation in 32 MHz anode damper resistor, due to the higher Pre-driver bandwidth coupled with the higher (30dB)FB, occurs when running at full power at 10 MHz. Also the Driver is pushed to near saturation by the higher (6dB) Local FB under these conditions. In spite of the addition in the RF input of a 10 MHz 3rd harmonic filter it was necessary to reduce the overall LG to 26dB when testing the modified amplifier in the ring. At 26dB FB the 'Herfurth' surface amplifier only delivers 25W at maximum gap voltage and the 3rd harmonic content is below –30dBC.

When testing the modified amplifier at full power in the ring the Mosfet Transistors failed several times. Examination of the gap FB capacitor showed breakdown of the 'N' coax connector of the 3ns cable. The FB capacitor system has been modified but at the time of writing no testing in the ring has been possible.

A bias correction circuit for compensation of the transistor working point due to radiation effects has been added to the test amplifier, but the early failures prevented a full evaluation of this idea.

For a fuller description of the development of the amplifier modified with transistors see the Diploma Thesis of Anton Labanc[3] at G:Users/g/grier/Public/Thesis.

6 CONCLUSIONS

The PS 10 MHz cavity impedance can be reduced at resonance by up to 10dB by increasing the power amplifier fast feedback loop gain. If a 10dB decrease in impedance is required without further increase in the input drive power, extensive modifications are necessary. The Pre-driver gain x bandwidth product must be increased by replacement of the tube by an RF Mosfet transistor.

Along with a rebuild of the grid resonator to reduce stray resonances, the increased gain can be used partly to reduce the loop group delay by increased local feedback.

The gap impedance can thus be reduced from the present 320ohms to 150ohms at 26dB loop gain and 112 ohms with 30dB. At 30dB gain difficulties have arisen with saturation of the Driver stage by harmonics from the drive amplifier and even at reduced gain (26dB) failure of the input transistor via voltage breakdowns occurred when testing in the ring More testing would be necessary before any decision could be made to modify amplifiers in the ring in this way.

An increase of 6dB from around 20dB to 26dB can be achieved with minor modifications provided a small decrease in gain margin and the increase in measured gap impedance (440ohms to 580ohms) at 32 MHz is accepted. The input drive power for maximum gap voltage at 10 MHz will be increased from 25W to near 80W under these conditions.

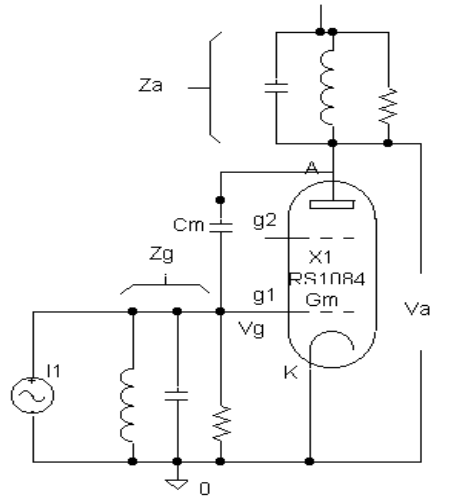
Four amplifiers incorporating these minor modifications have been working well in the PS ring during 2001, the only apparent difficulty being the 80 W required from the Herefurth input driver amplifiers. In 2002 all 11 amplifiers in the ring will have been modified to have 26dB nominal loop gain. During the 2002/2003 shutdown the Herfurth amplifiers will need modifying to reduce the 2nd and 3rd harmonics at 100W output.

Some of the ideas used on the 30dB transistorised version could, in the long term, be further investigated and exploited to improve the present amplifier.

a. Rebuilding of the grid resonator would decrease the stray winding capacitance thus reducing slightly the circuit Q and eliminating the strong 35/40 MHz resonance in the Driver stage.
b. A Final grid neutralisation circuit may work better in this amplifier because of the lower cutoff frequency of the Pre-driver stage.
c. Taking the overall feedback from the gap would improve the gain margin at 32 MHz by 15dB and lower the gap impedance at this frequency.

ANNEX 1

Miller effect in the Final stage



In the simplified circuit shown, Cm is the Cag1 internal to the tube. I1 is the anode current of the Driver tubes.
The Miller effect in such a stage with resonant grid and anode circuits is more complex than in the classical case of a wideband amplifier.

The mechanism can be understood by calculating the anode and grid signals as those of a simple amplifier where A is high (~100) but the loop gain Aβ is small (~1.2) with I1 as current source and Va as the output.

$$Cm := 2\cdot 10^{-12} \quad i := \sqrt{-1} \quad Is := 0.5 \quad \omega r1 := 2\cdot\pi\cdot 4.953\cdot 10^{6} \quad Gm := 40\cdot 10^{-3}$$

$$\omega r2 := 2\cdot\pi\cdot 5.1\cdot 10^{6} \quad Qa := 67 \quad Ra := 2500 \quad Qg := 3 \quad Rg := 200$$

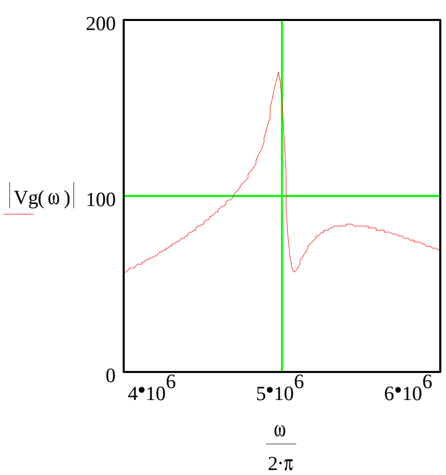
$$\omega := 2\cdot\pi\cdot 4\cdot 10^{6}, 2\cdot\pi\cdot 4.01\cdot 10^{6} .. 2\cdot\pi\cdot 6\cdot 10^{6}$$

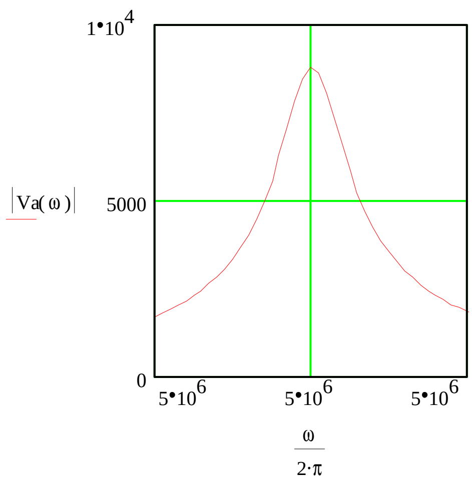
$$Za(\omega) := \frac{Ra}{1 + i\cdot Qa\cdot\left(\frac{\omega}{\omega r1} - \frac{\omega r1}{\omega}\right)} \qquad Zf(\omega) := \frac{1}{\omega\cdot Cm\cdot i}$$

$$A(\omega) := -Gm\cdot Za(\omega)$$

$$Zg(\omega) := \frac{Rg}{1 + i\cdot Qg\cdot\left(\frac{\omega}{\omega r2} - \frac{\omega r2}{\omega}\right)} \qquad \beta(\omega) := \frac{Zg(\omega)}{Zf(\omega) + Zg(\omega)}$$

$$Va(\omega) := \frac{Zf(\omega)}{\left(\frac{1}{\beta(\omega)\cdot A(\omega)}\right) + 1}\cdot Is \qquad Vg(\omega) := \left(Is + \frac{Va(\omega)}{Zf(\omega)}\right)\cdot Zg(\omega)$$





Close to the anode (bare cavity) resonance the amplifier stage operates with positive (regenerative) feedback when the source frequency lies below the nominal f_0, and with negative feedback when it is above. The peak of the anode resonance is thus shifted both in amplitude and, since the amplifier output impedance is complex, in frequency. Both the anode and grid resonators must then be retuned. If the grid resonator Q is not sufficiently low or if the stage Miller loop gain is too high neutralisation may be necessary.

ANNEX 2

Increasing the G x BW product with a transformer

Given the constraint of increasing the amplifier gain, without decreasing the bandwidth and increasing loop phase shift, is not an easy task. Adding an extra stage would increase the forward gain but only at the expense of the added stage phase shift. Increasing the gain of the existing stages without increasing the phase shift would only be possible if the stages were not already optimized.

Pulsing the Final tubes to a higher bias level so as to run in class A without excessive power dissipation would be a possibility, but this complicates the operation and makes measurements with the Network Analyzer laborious and so should only be adopted in the last resort. A second possibility is to add a matching transformer of step down ratio n between Pre-Driver and Driver so as to increase the gain x bandwidth product as shown below. L1 is an HT isolating inductor, Cs the stray capacitance of L1 plus Cag2 of X1. Cgt includes the Cgk and Cg1g2 of X2.

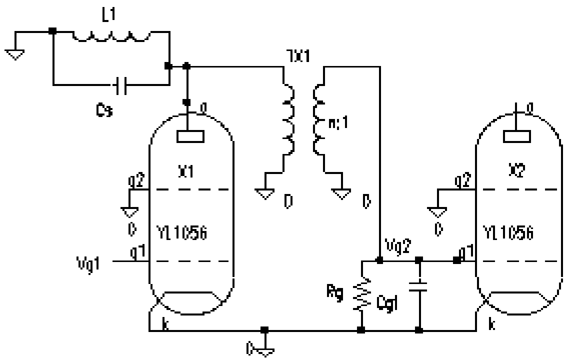


Fig. 1(annex) : Simplified equivalent circuit of Pre-Driver/Driver stages

In Fig. 1, between Vg1 and Vg2,

$$G \times BW = \frac{gm \cdot n \cdot Rg}{2\pi \cdot \left(Cgt + n^2 \cdot Cs\right) \cdot Rg} = \frac{gm \cdot n}{2\pi \cdot \left(Cgt + n^2 \cdot Cs\right)}.$$

If we impose a worst case of $Cs = \frac{Cgt}{n^2}$ then $G \times BW = \frac{gm \cdot n}{4\pi \cdot Cgt}$. For a transmission line transformer with integer n, for n=2 we there is no improvement unless Cs<Cgt/4. This appeared to be the case since the Cag2 of a YL1056 is 9pF as compared to 270pF at the grids of the two Driver tubes, but the additional anode strays along with capacity of L1 added up to a total of 50pF (Cgt/5.5). In practice a small improvement in bandwidth was obtained and very little improvement in phase lag over that already obtained with the compensating inductor L6. Some resonances between the transformer and the coupling capacitor (not shown) are difficult to suppress.

References

[1] A. Blas, R. Garoby, “Design and Operational Results of a “One-Turn-Delay Feedback” for Beam Loading Compensation of the CERN PS Ferrite Cavities”, CERN/PS 91-16 (RF).

[2] M. Paoluzzi, private communication.

[3] A. Labanc, Technical Student from Bratislava Univ. Thesis project 2000/2001.

Acknowledgements

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