THE RF COUPLING SYSTEM OF THE VINCY CYCLOTRON

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ABSTRACT

The main parameters of the VINCY cyclotron RF system are presented. A possibility to tune the λ/2 primary resonator only by action of a variable capacitor is discussed. A design procedure based on transformer equations has been verified by model measurements.

1. INTRODUCTION

The main parameters of the VINCY Cyclotron have been already presented[1] (k_b=150, both physical and medical purposes,...). In this isochronous cyclotron, the acceleration of ions on the first, second and fourth harmonic is accomplished by two dees settled in two opposite magnet valleys. The resonator type is implied by the basic design constraints. The use of the magnet based on the modification of the CEVIL magnet from Orsay, France, implies application of horisontal λ/4 type resonator cavities. The design of the cavities is based on the use of the prototype cavities of the LNS Catania Superconducting Cyclotron [2]. A short circuiting piston is used to tune the cavities thruoghout the frequency range. The main parameters of the RF system are:

Table 1.

Accelerating dees	2 x 30°
Min. RF gap	10 mm
Frequency range	17-31 MHz
Piston excursion	2 m

In order to avoid complex matching system, each power amplifier is directly coupled to the corresponding resonator cavity, thus enabling the use of only one variable vacuum capacitor as a matching element. Each amplifier makes a part of a λ/2 primary resonator, inductively coupled with the cavity. The λ/2 resonators were chosen in order to locate the power amplifiers in a low stray magnetic field zone (fig 1.) where the power tubes could be efficiently screened.

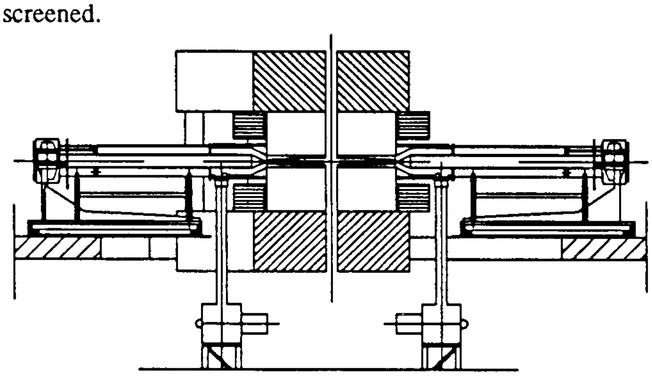

Fig. 1. The section layout of the VINCY cyclotron.

Having chosen inductive coupling, one has the choice between two solutions.

a) *Tuned primary, tuned secondary.* The obvious drawbacks are first a need of moving the mechanical coupling loop inside the vacuum to adjust the primary impedance level and second, a need for a fast electronic device to protect the power tube in case of arc sparking at the dee. The advantages are that the isolated tuning frequencies of both resonators coincide with the accelerating frequency.

b) *Detuned primary.* The coupling loop is fixed, the impedance level is obtained by detuning considerably the primary resonator. The reactive energy required by this resonator can be delivered only by the secondary one which, in turn, is also detuned, in the opposite sense, though by a much smaller amount. It follows that the isolated resonant frequencies of the primary and secondary, ω_{n1}, ω_{n2}, must be known beforehand. Any subsequent adjustment of the primary impedance (which is easily obtained with the variable condenser on the primary) requires an adjustment on the secondary as well. If this cannot be obtained by the fine tuning paddle only, the piston position should be corrected - a risky action at full power operation.

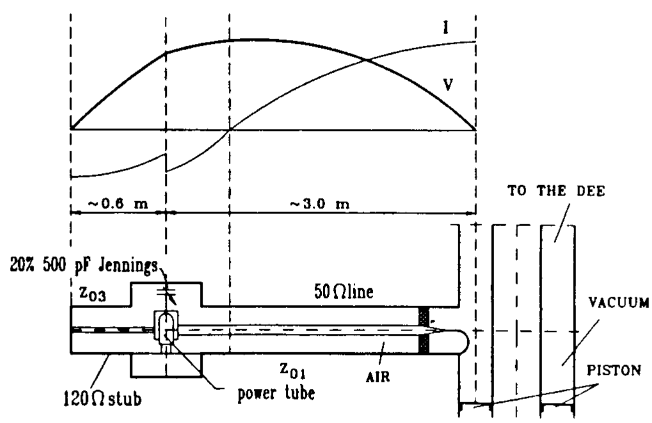

Fig. 2. How the resonators are coupled.

An other important fact is that the coupling loop size is noticeably larger than in a) and that there is a limitation due to mechanical constraints (sparking danger between the loop and the secondary stem).

For these reasons and in order to check the feasibility of the arrangement over the whole RF band, an analysis has been developed which shows the mechanism of the coupling phenomena and, by means of simple expressions, calculates the characteristic and the criticity of the components involved.

2. THE TRANSFORMER EQUATIONS

Instead of the standard approach by using Kirchhoff equations, we have preferred one which highlights the feedback action between primary and secondary resonators as a function or the coupling coefficient k. This leads to the primary impedance as seen by the power tube:

$$Z_1(S) = \frac{\omega_n Z_{n1v}}{2} \frac{S+\sigma_2}{(S+\sigma_1)(S+\sigma_2)+\Omega^2} \qquad [2.1]$$

As explained[4], the expression is approximate and it is acceptable for bandwidths of the order less than 10% and Q's of the order of 1000. Z_{nv} is a proportionality constant between top dee voltage V_1 squared and energy content E:

$$Z_{n1v} = \frac{V_1^2}{2\omega_n E} \qquad [2.2]$$

σ is the 3-db half-bandwidth : $\sigma = \omega_n/2Q$ and

$$S = s - j\omega_n \qquad [2.3]$$

is the complex displacement from the nominal frequency $j\omega_n$. Though not absolutely necessary, reduced variable S is used (narrow band approximation)[4], which cuts by half the degree of the polynomials at the price of systematic errors. It is found that, as in our case, when the Q's are of the order of several thousands and the relative bandwidths less than 10%, the errors are tolerable.

Ω is a frequency related to the coupling coefficient k

$$k = \frac{\omega_n}{\sqrt{Z_{n2i\text{-}loop} Z_{n1i\text{-}loop}}} M$$

$$\Omega = \omega_n k/2 \ ; \qquad M = \frac{\mu_o A}{2\pi r} \qquad [2.4]$$

(r = distance of the coupling loop gravity centre from the linear axis; A the loop surface).

An other constant, which appears in the coupling coefficient k, is Z_{ni} defined via the internal circulating current i:

$$Z_{n1i} = \frac{2\omega_n E}{i^2} \qquad [2.5]$$

Both Z_n's have the dimension of an impedance and are functions of the point x where the resonator is measured.

The advantage of using the Z_n's instead of the equivalent inductances and capacitances is that the former are simple figures of the order of 50 Ω, independent of the frequency. For our purpose, only the dee top and the loop position are interesting. For the secondary, which is a non-uniform transmission line, all $Z_{nv}(x)$ and only $Z_{ni\text{-}short}$ can be measured with the perturbation method or calculated by means of the section method[5] and also by means of modern codes[2]; $Z_{ni\text{-}loop}$ can only be obtained by calculation.

For the primary, if the power amplifier can be considered as a lumped capacitance tuning a φ long transmission line Z_o, we find :

$$\begin{cases} Z_{nv\text{-}anode} = 4Z_o \sin^2\varphi / (\sin 2\varphi + 2\varphi) \\ Z_{nv\text{-}loop} = Z_o (\sin 2\varphi + 2\varphi)/4 \end{cases} \qquad [2.6]$$

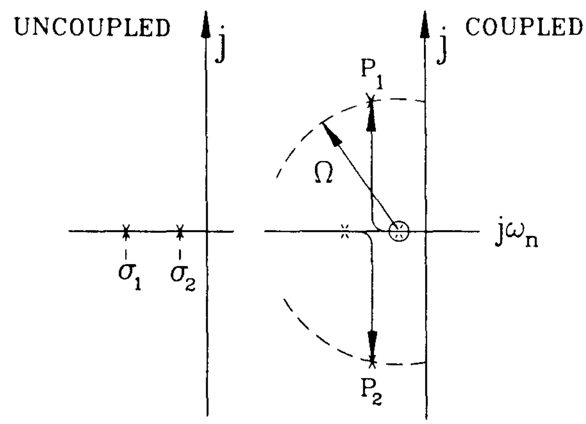

Fig 3. The singularities of the tuned primary arrangement

3. THE TUNED PRIMARY CASE

The denominator of Eq. [2.1] could be factorized as:

$$Z_1 = \frac{\omega_n Z_{n1v}}{2} \frac{S+\sigma_2}{(S-P_1)(S-P_2)} \qquad [3.1]$$

where $$P_{1,2} = -\frac{\sigma_1+\sigma_2}{2} \pm j\sqrt{\Omega^2 - (\frac{\sigma_2-\sigma_1}{2})^2} \qquad [3.2]$$

therefore, the locus of the coupled poles (Fig. 3) show that by increasing k, the poles move vertically still staying on a circle centered at $j\omega_n - \sigma_2$ with radius Ω. A zero appears where there was previously the uncoupled pole at $-\sigma_2$. The exact Kirchhoff solution would indicate a zero at $j\omega_n/\sqrt{1-k^2}$ and a pair of poles at $j\omega_n/\sqrt{1-k}$ and $j\omega_n/\sqrt{1+k}$. The lack of geometrical symmetry is due to the narrow band approximation. A plot of amplitude and phase is shown in Fig. 4. The impedance is real at ±jΩ and at centre. Letting S = 0 in Eq. [2.1], the resulting $Z_1(0)$ is the parallel between the primary R_p : $R_p = \omega_n Z_{niv}/2\sigma$ and the reflected secondary impedance Z_R :

$$Z_R(0) = \frac{\omega_n Z_{n1v}}{2} \frac{\sigma_2}{\Omega^2} \qquad [3.3]$$

This expression, together with the energy equation, leads to the desired transforming ratio V_2/V_1 :

$$\tau = \sqrt{\frac{Z_{n2v}}{Z_{n1v}}\frac{\Omega}{\sigma}} \qquad [3.4]$$

Both t and Z_R depend on Ω, and so finally on the cross section A of the loop.

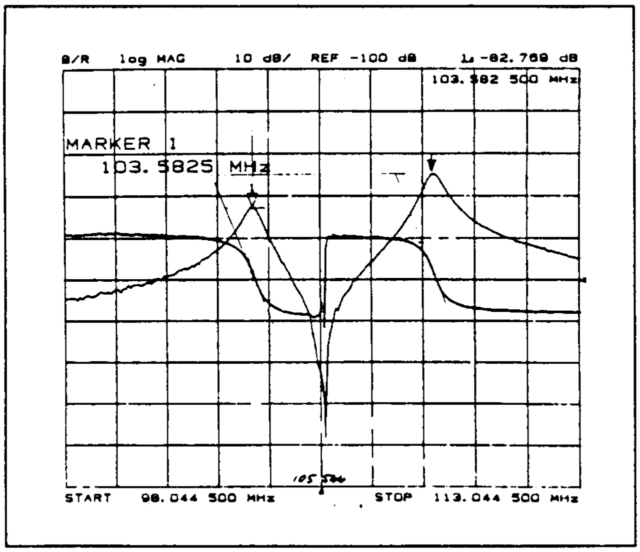

Fig. 4. The phase and amplitude function.

4. DETUNING THE PRIMARY

Because the poles were located in the S plane at $j0 - \sigma$, the terms $S + \sigma$ appeared in Eq. [2.1]. If the primary had been resonant at $\omega_n - \delta$ (detuned), then the pole, still in the S plane would have been located at $-j\delta - \sigma$ and a term $S + \sigma + j\delta$ would have appeared at the denominator instead. To facilitate the discussion, the primary pole is supposed to be affected by a real part $\sigma_1 = \sigma_2 = \sigma$. In the obvious case where $\sigma_1 > \sigma_2$ a resistance $R_p = \omega_n Z_{n1}/2(\sigma_1 - \sigma_2)$, which does not participate to the energy transfer, will be considered to be in parallel to the primary. For the same reason, the power tube internal resistance has to be neglected. It should, however, be taken into account in those cases where the beam load transfers energy to the accelerating secondary, because instabilities could occur[4].

Since the primary resonator is tuned at $\omega_n - \delta$, the coupled primary impedance is

$$Z_1(S) = (\omega_n - \delta)Z_{n1v}\frac{S+\sigma}{(S+\sigma)(S+\sigma+j\delta)+\Omega^2} \qquad [4.1]$$

with roots : $$P_{1,2} = -\left(\sigma + j\frac{\delta}{2}\right) \pm j\sqrt{\Omega^2 + \delta^2/4} \qquad [4.2]$$

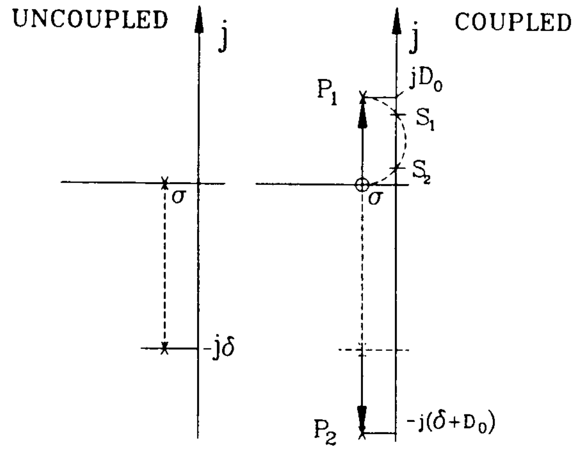

Fig 5. The singularities of the detuned primary arrangement

Again, there are three frequencies where the impedance is real : close to P_1, close to S = 0 and at $P_2 = -j(d + D_0)$. From figure 5 it appears that:

$$D_0 = \sqrt{\Omega^2 + \delta^2/4} - \delta/2 \quad \Rightarrow \quad D_0(D_0 + \delta) = \Omega^2 \qquad [4.3]$$

The frequency at P_2 is the primary mode, non interesting since no significant energy is transmitted to the secondary.

The two other frequencies with phase zero are found at the intersection between the circle diameter D_0 with the j axis. Generally, S_2 leads to excessively low impedances whereas

$$S_1 = D_0 - \sigma^2/D_0 \qquad [4.4]$$

is the useful one, provided of course that

$D_0 >> 2\sigma$ (say $D_0 = N\sigma$)

It is found that for N = 3 the upper phase margin is less than 20° and it is doubtful that the automatic tuning system could work satisfactorily in the presence of a noise or transients.

The relation between D_0 and δ can be obtained from energy considerations : there are two resonators, the primary tuned at $-j\delta$ and the secondary at j0; both are excited at $j(D_0 - \sigma^2/D_0)$. No reactive power is delivered by the amplifier at $S = S_1$. The energy is balanced between the two resonators since at S_1 there is a 180° phase shift. Therefore, it must be :

$$\frac{V_2^2}{2}\mathrm{Im}(Y_2(S1)) = \frac{V_1^2}{2}\mathrm{Im}(Y_1(S1)) \qquad [4.5]$$

which leads to :

$$\frac{V_2^2}{\omega_n Z_{n2v}}\left(D_0 - \frac{\sigma^2}{D_0}\right) = \frac{V_1^2}{(\omega_n - \delta)Z_{n1v}}\left(D_0 + \delta - \frac{\sigma^2}{D_0}\right) \qquad [4.6]$$

and finally: $$\delta\left(\frac{1}{D_0 - \sigma^2/D_0} + \frac{\tau^2}{\omega_n}\frac{Z_{n1v}}{Z_{n2v}}\right) = \tau^2 \frac{Z_{n1v}}{Z_{n2v}} \quad [4.7]$$

5. THE VINCY ARRANGEMENT

In our primary resonator shown schematically in Fig. 2, the amplifier is located in the second half of the resonant transmission line Z_{01}. The characteristic impedance Z_{03} of the tuning stub and its length x_3 can be negotiated to keep the amplifier sufficiently far from the median plane and compatible with the physical dimension of the power cabinet.

To calculate Z_{nv} and Z_{ni} for this arrangement, from Eqs. [2.2] and [2.5]:

$$\frac{1}{Z_{nv\text{-anode}}} = \frac{1}{Z_{nv\text{-line1}}} + \frac{1}{Z_{nv\text{-line3}}} \quad [5.1]$$

Since $$I_{loop\text{-line1}} = V_{anode} / Z_{01} \sin\varphi_1 \quad [5.2]$$

whereas the energy is: $$E = V^2_{anode} / (2Z_{nv\text{-anode}}) \quad [5.3]$$

gives: $$Z_{ni1\text{-loop}} = Z^2_{01} \sin^2\varphi_1 / Z_{nv2\text{-anode}} \quad [5.4]$$

Once the Q's are known, the Z_{nv}'s give the shunt resistances

$$R_p = Z_{nv}\, Q \quad [5.5]$$

and the power in function of the applied voltages.

a) *Primary*. It has been shown in Figs. 1 and 2. The power tube is a SIEMENS type RS1084CJ tetrode in grounded grid configuration. The anode d.c. voltage is 10 kV, then V_{rf1} = 8.5 kV. As the maximum dee voltage will be V_{rf2}= 100 kV, independent of the frequency, then the transforming ratio is $\tau = 11.7$

A wooden model of the cabinet with tuning lines and coupling loop indicate that the primary resonator can be considered as a x_1 = 2.77 m long 50 Ω line terminated at its open end with x_3 = 0.6 m long 120 Ω line in parallel with a variable vacuum capacitor variable between 20 and 423 pF at 31 and 17 MHz respectively.

Then, from Eq. [2.5, 2.6] one gets (Table 2.):

Table 2.

f (MHz)	31	17
$Z_{n1v\text{-anode}}$ (ohm)	26.26	16.73
$Z_{n1i\text{-loop}}$ (ohm)	90.17	104.22

b) *Secondary* The dimensions of the secondary resonator are such that r = 0.225 m. The other parameters, calculated and measured[2] are (Table 3.) :

Table 3.

f (MHz)	31	17
Z_{n2v} (ohm)	42.25	52.98
$Z_{n2i\text{-loop}}$(ohm)	31.5	58.73
σ(kHz)	2.25	1.42
Q_2	6880	5980
R_{p2} (ohm)	291	317
W_2(kW)	17.2	15.7

c) *With the tuned primary* coupling, one obtains

Table 4.

f (MHz)	31	17
Ω (kHz)	22.9	9.36
A (cm^2)	5.42	11.0

which requires a mechanical displacement or rotation in vacuum of the coupling loop.

d) *With detuned primary* one starts at low frequency choosing D_0 sufficiently higher, though not excessively, then σ. From the phase plot it has been adopted N = 7. From Eq. [4.4] one gets:

D_0 = 7 x 1.42 = 9.94 kHz.

From Eq. [4.5] we get : δ = 213 kHz; S_2 = 9.73 kHz.

From Eq. [4.2] and [2.9]: Ω = 47 kHz and A = 50,7 cm^2.

At high frequencies, introducing the same value D_0 we get A = 19.5 cm^2. By iteration, it is found that D_0 = 22.5kHz leads to the previous value for A :

δ = 2030 kHz; S_2 = 25.25 kHz; Ω = 228.9 kHz; A = 50.7 cm^2.

6. CONCLUSION

It has to be observed that tuning the dee over the whole band as well as changing the transforming ratio τ can be performed by a unique action on the variable vacuum capacitor provided the dee fine tuning facility succeeds to retune from S_2 to zero i.e. at ω_n. Otherwise, the piston of the dee line has to be displaced.

7. REFERENCES

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