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Power Supply Ripple Study at the SPS

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Abstract

Power supply ripple of the main dipoles and quadrupoles may further limit the dynamic aperture of large proton colliders like the LHC in the presence of the non-linear field errors of the superconducting magnets. The SPS, the future injector of the LHC, is used to evaluate the effects of power supply ripple. In this report we demonstrate how the ripple spectrum measured at the power supplies of the different SPS magnet chains can be translated into an expected spectrum of tune ripple seen by the beam. To this end the transmission line characteristics of various single elements and magnet chains have been measured and compared to theoretical models. The damping effect of eddy currents induced in the vacuum chamber by the rippled main fields has been measured and compared to a preliminary model. The resulting tune spectrum is then compared with the actual tune ripple which has been measured with the Schottky system. These investigations will allow to specify the quality of the LHC main power supplies.

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1 Introduction

Conventional power supplies of the guiding and focusing magnets always introduce unwanted AC components of 50Hz (60Hz in the US) and multiples of it, in particular a large 600Hz component due to the 12-pole converter of the power supply rectification system. The delicate collider operation of the $SP\bar{P}S$ is an example of how important this power supply ripple can be. For a satisfactory performance as a collider it was necessary to reduce the power supply ripple which caused repeated crossing of dangerous beam-beam resonances [1]. This has been achieved by carefully adjusting the power supplies including a reduction of the 50Hz component with active filters.

It is also well known by now that these low ripple frequencies can strongly reduce the dynamic aperture ¹ in the presence of strong non-linearities. A thorough study [2] has been launched to quantify this combined effect of non-linearities and tune ripple caused by power supply ripple.

Of course all these studies are motivated by the design of the LHC, the new proton collider proposed to be installed in the LEP tunnel. The demands for the highest possible fields make superconducting magnets mandatory which do, however, lead to large unavoidable multipolar errors. Given these errors we still have to find the minimal level at which tune ripple starts to limit the dynamic aperture of the LHC. To avoid the problem it may be sufficient to use a more modern technology for the main power supplies, like the switch mode power converters [3] which do not introduce any appreciable ripple below the chopping frequency ($> 10\text{kHz}$). However, this is a fairly recent technology and at present it is not completely clear if the LHC magnets with their large currents ($\leq 11.5\text{kA}$) can be powered in this way and how costs compare to that of a more conventional system.

In this article we will show to which extent we understand the behavior of the main dipole chains of the SPS as transmission lines. The knowledge of their transfer functions or impedances can be used together with a measurement of the voltage ripple at the power supplies to derive the tune ripple seen by the beam. Of course, all effects have to be considered which may alter the transfer functions such as the damping effect of the vacuum chamber. The resulting tune ripple spectrum will then be compared with a direct measurement of the tune ripple of the beam using the Schottky system [4].

In a future study we will use our understanding gained from these SPS measurements to estimate an electric equivalent of the LHC magnets chains which will allow us to evaluate tune ripple spectra. These variations of the tunes will be studied in tracking with the goal to specify the above mentioned minimal quality requirement of the main power supplies of the LHC.

¹The dynamic aperture can be defined as that distance to the closed orbit at which the injected particles will just barely stay in the accelerator.

2 Theoretical Background

2.1 Four-Poles

To understand the theory of transmission lines it is convenient to start with four-poles. The four-poles of interest are passive (no internal energy sources) and linear that is to say they are made of linear components which are resistors, capacitors or inductors. It can be shown [5] that these four-poles have the property of being reciprocal: that means that for a four-pole with an input voltage at one end and a short circuit at the other end, the current in the short circuit is the same independent of which end is used for the input voltage. This reduces the 8 free parameters (complex voltage and current at either end) to 6 so that such a four-pole can be represented by two equivalent forms: a triangular shape like in Fig. 1a or a star-like shape as in 1b. In the following we will use type a.

Figure 1: Definition and Examples for Four-Poles

Moreover in most cases we will deal with symmetric four-poles which have $Z1$ equal to $Z3$ (see Fig. 1a). A four-pole can be seen as a matrix A (4 by 4) which transforms the right-hand voltage U_2 and current I_2 into the left-hand values U_1 and I_1 respectively (all values are complex and are treated as two dimensional vectors):

$$(1) \quad \begin{pmatrix} U_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{pmatrix} \times \begin{pmatrix} U_2 \\ -I_2 \end{pmatrix}.$$

The matrix components can be expressed in the impedances of Fig. 1a:

$$(2) \quad A_{11} = 1 + \frac{Z2}{Z3}, \quad A_{22} = 1 + \frac{Z2}{Z1}, \quad A_{12} = Z2, \quad A_{21} = \frac{1}{Z1} + \frac{1}{Z3} + \frac{Z2}{Z1 * Z3}.$$

In the symmetric case ($Z1 = Z3$) the diagonal elements become equal:

$$(3) \quad A_{11} = A_{22} = 1 + \frac{Z2}{Z1}, \quad A_{12} = Z2, \quad A_{21} = \frac{2}{Z1} + \frac{Z2}{Z1^2}.$$

In Fig. 1 we have also shown four simple examples of four-poles. The first (denoted by 1) are 2 wires with $Z1 = Z3 = \infty$. This is a symmetric four-pole, the identity matrix:

$$(4) \quad A(1) = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}.$$

Example 2 is also symmetric:

$$(5) \quad A(2) = \begin{pmatrix} 1 & Z2 \\ 0 & 1 \end{pmatrix}.$$

The example 3 is not symmetric:

$$(6) \quad A(3) = \begin{pmatrix} \infty & 0 \\ \infty & 1 \end{pmatrix}$$

The infinite values imply that U_2 has to be zero (the circuit is a short) and I_1 is simply equal to $-I_2$. In example 4 we have combined 2 and 3:

$$(7) \quad A(4) = \begin{pmatrix} \infty & Z2 \\ \infty & 1 \end{pmatrix}.$$

2.2 Transmission Line

A short transmission line can be described as in Fig. 2 with the electric components taken per unit length (denoted by the dash). R' , L' and C' stand for resistor, inductor and capacitor respectively, while G' is the inverse of the resistor which is in parallel with the capacitor C' .

Figure 2: Electric Equivalent of short Transmission Line

The figure leads intuitively to a differential equation which has the following solution for the voltage U and the current I :

$$Ue^{i\omega t} = a_1 e^{-\alpha x} e^{i(\omega t - \beta x)} + a_2 e^{\alpha x} e^{i(\omega t + \beta x)}$$

$$(8) \quad Ie^{i\omega t} = \frac{a_1}{Z_o} e^{-\alpha x} e^{i(\omega t - \beta x)} - \frac{a_2}{Z_o} e^{\alpha x} e^{i(\omega t + \beta x)},$$

where Z_o is the wave impedance, a_1 and a_2 are two complex constants yet to be determined, α is the damping and β is related to the phase velocity v via:

$$(9) \quad v = \frac{\omega}{\beta}.$$

The complex damping γ combines the damping α and β :

$$(10) \quad \gamma = \alpha + i\beta.$$

The parameters Z_o , α and β can be expressed in the electric components of Fig. 2:

$$Z_o = \sqrt{\frac{R'^2 + \omega^2 L'^2}{G'^2 + \omega^2 C'^2}} \quad \alpha = \frac{\omega(R'C' + L'G')}{\beta}$$

$$(11) \quad \beta = \sqrt{0.5(-R'G' + \omega^2 L'C') + 0.5\sqrt{(R'^2 + \omega^2 L'^2)(G'^2 + \omega^2 C'^2)}}.$$

For $(\omega \gg 1)$ α becomes independent of ω and reduces to:

$$(12) \quad \alpha = \frac{R'C' + L'G'}{L'C'}.$$

For negligible damping ($R' \ll 1$ & $G' \ll 1$) β reads:

$$(13) \quad \beta = \omega\sqrt{L'C'}.$$

The solution of eq. 8 can be interpreted as an incoming wave with amplitude a_1 and a reflected wave with amplitude a_2 . An example is depicted in Fig. 3 showing these and the resulting wave.

Figure 3: Example of Propagation of Waves in finite Transmission Lines
(signal versus normalized length)

The decay of the incoming wave, which is e^{-ax} , is shown as well.

Using the initial conditions we can now solve for a_1 and a_2 and we arrive at:

$$(14) \quad \begin{pmatrix} U_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} \cosh(\gamma l) & Z_o \sinh(\gamma l) \\ \frac{1}{Z_o} \sinh(\gamma l) & \cosh(\gamma l) \end{pmatrix} \times \begin{pmatrix} U_2 \\ -I_2 \end{pmatrix},$$

in which l denotes the length of the transmission line. As the diagonal elements are equal, a transmission line can be seen as a symmetric four-pole. Combining a chain of n identical elements to one long transmission line can be formally achieved by simply replacing γl by $n \times \gamma l$. The wave impedance Z_o does not change in this operation, a termination with Z_o on one end eliminates the reflected wave and the same impedance is measured at the other end of the transmission line.

The impedances Z_{op} and Z_{cl} at one end, the other end being open or closed respectively, allow to derive Z_o and γ :

$$(15) \quad \begin{aligned} Z_{op} &= Z_o \frac{1}{\tanh(n \times \gamma l)} & Z_{cl} &= Z_o \tanh(n \times \gamma l) \\ \Rightarrow Z_o &= \sqrt{Z_{op} Z_{cl}} & \tanh(n \times \gamma l) &= \sqrt{\frac{Z_{cl}}{Z_{op}}}. \end{aligned}$$

Another important feature of the transmission lines is the existence of resonances. For convenience we study the case with very small damping ($\alpha \approx 0$). Then the $\tanh$ of Z_{op} and Z_{cl} in eq. 15 can be replaced by $\tan$ which gives rise to infinite values whenever the argument becomes a multiple of $\frac{\pi}{2}$. The resonance frequency f_r is therefore simply:

$$(16) \quad \beta l = \frac{\pi}{2} \quad \Rightarrow \quad f_r = \frac{1}{4\sqrt{LC}}.$$

Notice that since we multiply with $n \times l$ the values L and C are here the inductor and capacitor of the total transmission line (no dashes). At the resonance frequency the impedances at the ends Z_{E1} , Z_{E2} are related as follows:

$$(17) \quad Z_{E1} = \frac{Z_o^2}{Z_{E2}}.$$

This inversion means for instance that a short circuit on one end will enforce an infinite impedance on the other side. Of course, an impedance Z_o at one end leads to Z_o at the other end.

When the damping is included the impedances Z_{op} and Z_{cl} also show peaks at the same frequency values as without damping, however they are damped to finite values which decrease exponentially with frequency.

2.3 Shielding due to Vacuum Chamber

Besides the damping of the transmission lines there is an additional damping due to the vacuum chamber. The alternating fields due to power supply ripple induce eddy currents in the vacuum chamber which act against the exciting fields so as to shield the field inside the vacuum chamber.

The analytic treatments [6] of eddy currents usually have to be restricted to circular chambers. In more modern approaches numerical tools start to be included [7]. In the SPS, however, we have more complex forms of the vacuum chambers (e.g. the right part of Fig. 4). Moreover we have to include the iron of the magnets in the treatment. The results of previous studies are therefore of little help in our practical application. A work by M. Gyr [11] looked therefore very attractive at first sight. In his study the vacuum chamber is replaced by segments which carry the proper induced currents. These segments can build up any form of vacuum chambers. In addition the effect of iron of magnets is taken into account by considering mirror images of the currents in the chamber segments. However, the treatment stays simple only for constantly ramped fields. Once sinusoidal fields are involved, self-inductance and the mutual inductance between each pair of the current segments has to be included which makes the method impractical. The only way to find reliable results which can be compared with our measurements seems to be a numerical field calculation inside the real vacuum chamber and the magnetic poles of the magnets.

For our purposes here we follow a paper of R. Shafer [8] which also treats a simplified structure of a circular $\cos(\phi)$ current distribution at radius a (see left part of Fig. 4) and a circular vacuum chamber at radius b . It arrives at an electric equivalent of the vacuum chamber which is quite instructive and which we use for fitting our measurements.

Figure 4: Scheme of Magnet with Vacuum Chamber

Left part shows the theoretical model and right part the measured MBB-s. The shape of the vacuum chamber in the MBB-s is slightly simplified – values are in units of [mm].

Shafer calculates the vector potential of the structure described above in which he allows for any multipolar field of order n at a radius a . For simplicity we will restrict ourselves to a dipole field ($n = 1$). Starting with a undisturbed magnetic field:

$$(18) \quad B(t) = B_o \sin(\omega t),$$

we arrive at a field with vacuum chamber:

$$(19) \quad B(t) = B_o \frac{1}{\sqrt{1 + \omega^2 \tau^2}} \sin(\omega t - \arctan(\omega \tau)).$$

The fields are therefore shielded by the factor $\frac{1}{\sqrt{1 + \omega^2 \tau^2}}$ and retarded by $\arctan(\omega \tau)$ where τ is the time constant defined as:

$$(20) \quad \tau = \frac{\mu_o b d}{2 n \rho},$$

where μ_o is the magnetic field constant, d is the thickness of the vacuum chamber and ρ is the resistivity in units of $[\frac{\Omega \times \text{mm}^2}{\text{m}}]$.

Calculating the total induced voltage we arrive at the impedance which represents the inductor of the magnet plus the vacuum chamber. The inductor becomes frequency dependent and there is also a frequency dependent resistive part:

$$(21) \quad L(\omega) = (1 - k)L + \frac{kL}{1 + (\omega\tau)^2}$$

and

$$(22) \quad R(\omega) = \frac{(\omega\tau)^2 R_s}{1 + (\omega\tau)^2}$$

As shown in Fig. 5 the inductor is replaced by two parts:

Figure 5: Electric Equivalent of Vacuum Chamber

$[(1 - k) * L]$ and $[k * L]$, the latter having in parallel a resistor R_s :

$$(23) \quad R_s = \frac{kL}{\tau}.$$

This can be interpreted in the following way: at very low frequency ($\omega kL \ll R_s$) $L(\omega)$ stays almost at the undisturbed value L , at some frequency ($\omega kL \approx R_s$) the inductor starts to be reduced and at large frequency ($\omega kL \gg R_s$) the inductor is constant but reduced by the factor $(1 - k)$. In the latter case, however, the field inside the vacuum chamber is almost completely shielded because the magnetic field energy of the kL part of the inductor is negligible compared to the energy lost in the resistive vacuum chamber which is represented by R_s .

3 Transmission Line Measurements at the SPS

3.1 Measurement Set-up

3.1.1 TRANSMISSION LINE MEASUREMENT

Similar transmission line measurements have been done at Fermilab, DESY and the SSCL [8, 9, 10]. As in the HERA study [9] we measure both the open and closed impedances (Z_{op} , Z_{cl} in eq. 15). Fig. 6 shows our set-up for this measurement.

Figure 6: Scheme of Transmission Line Measurement

For the frequency response analyzer we have used a Schlumberger 1250 which provides a sweeping sinusoidal output voltage from the sub-Hz level up to about 60kHz. The inputs are the voltage at the probe and the current measured in a variable pre-resistor R_p . The analyzed results (amplitude and phase of impedance as a function of frequency) are plotted with a HP plotter (unfortunately we did not manage to save the data in a file on a PC which would have allowed further processing). The intermediate amplifier (HP 467A power amplifier) has been usually set to a gain of one and is simply used to reduce the internal impedance of the voltage source.

It has to be stated that the resistor R_p is itself part of the transmission line (A_s) being studied. It is a four-pole of type 2 in Fig. 1. The actual transmission line matrix (A_{tm}) can be found by a simple matrix multiplication with the inverse of the matrix of a four-pole type 2:

$$(24) \quad A_{tm} = \begin{pmatrix} 1 & -R_p \\ 0 & 1 \end{pmatrix} \times A_s$$

3.1.2 SHIELDING MEASUREMENT

For the measurement of the shielding (as described in Fig. 7) the same set-up is used as in the above section (3.1.1), except that here only the induced current of the coil is recorded as a function of frequency.

Figure 7: Scheme of Measurement of Vacuum Chamber Shielding

The coil (internally named $D6$) is $6m$ long and consists of one winding on an insulated support, its resistance is 2.57Ω and its inductance is $30\mu H$. A transfer function measurement of this coil (as described above in section 3.1.1) has shown a linear behavior of the coil up to several 10kHz. This coil can be placed inside of the special MBB-s dipole magnet with and without a MBA type vacuum chamber. The result of the measurement of the induced voltage depends crucially on the choice of the pre-resistor R_p in Fig. 7. Simply applying Faraday's law of induction we arrive at an induced voltage U_I of:

$$(25) \quad U_I \propto \omega \times \cos(\omega t),$$

provided that

$$(26) \quad R_p \gg \omega L.$$

Once ωL approaches R_p the induced voltage is:

$$(27) \quad U_I \propto \cos(\omega t).$$

3.2 Electric Model of one Dipole Magnet

There are various sophisticated programs around to study transmission line effects, but we preferred to write ourselves a small program which is sufficient for our purpose. Of course, we have checked our results with these professional programs and have found perfect agreement. The best check, however, is to test the program with a real system. To this end we have put together an electric circuitry which represents (see the appendix, case 1 in Tab. 1) approximately one SPS dipole magnet (with the exception of the resistor RL in series with the inductor which was larger for technical reasons). Fig. 8 shows a schematic view of the circuitry: on the right hand side there is a switch which indicates that we are studying open and closed systems with respect to the yoke which is grounded. Moreover we can investigate the effect of a damping resistor R_d with the upper switch.

Figure 8: Model of Electric Equivalent of one Dipole Magnet

Fig. 9 summarizes the results of our comparison (0dB corresponds to the pre-resistor R_p). The solid lines (light for open and dark for closed circuits) show the measurements which agree very well with the dotted lines from the theoretical model (dark for open and light for closed circuits): in particular in the case of the closed system the curves are almost identical, the resonances are very well predicted (although the peaks of the theoretical curves are more pronounced). The internal damping is such that only one peak is visible. For a comparison we also show the theoretical curves (dashed lines) when the extra damping with $R_d = 10\Omega$ is switched on. No resonance structure is present in this case. Notice that both undamped cases approach the externally damped open circuit (light dashed line) in the limit of large frequencies.

Figure 9: Amplitude of Impedance of Electrical Equivalent of one Dipole

Figure 10: Phase of Impedance of Electrical Equivalent of one Dipole

3.3 Single Dipole Magnet and Shielding

Having found acceptable agreement with our theoretical model we will now treat the case of one dipole of the SPS, the MBB-s which is located in the magnet hall and accessible for measurements. The electric model is shown in Fig. 11.

Figure 11: Model of Dipole Magnet including Vacuum Chamber Shielding

Notice that the inductor L is already replaced by the electric equivalent of the vacuum chamber. The value of each component can be found in the appendix (Tab. 1, case 2). But at first we measure the magnet without the chamber. In Fig. 12 the measurement is seen as the black solid line. Our theoretical model (light dotted line) agrees very well with the measurement up to some frequency (2000Hz) and the resonance frequency is well predicted. To find a good fit we had, however, to add a mild damping ($R_d = 650\Omega$). This is an indication that the magnet structure (especially the iron) has a shielding effect as well. This may also explain the small deviation of the theoretical and experimental curve above 2000Hz. When the vacuum chamber is introduced (light solid line in Fig. 12) the scenario as described in section 2.3 can clearly be seen²: just below 200Hz the slope is about one, at 200Hz it decreases and above some 1000Hz it comes back to a slope of one (this is distorted at 5000Hz where the effect of the resonance starts to be visible). The fit (black dotted line) which uses a shielding resistor of $R_s = 6\Omega$ and a splitting factor $k = 0.5$ does agree surprisingly well. It is in fact even better than without the chamber which is another indication that the magnet structure has a shielding effect.

Well aware of the fact that the magnet structure and its interplay with the vacuum chamber is not included in Shafer's analysis we will nevertheless give an estimate of the expected value of R_s . Using eq. 23, the measured resistivity of $\rho = 0.8 \frac{\Omega \times \text{mm}^2}{\text{m}}$, the vacuum chamber thickness of $d = 2\text{mm}$ and varying the radius from $b = 21\text{mm}$ to $b = 76\text{mm}$ (the shape of the vacuum chamber is depicted in the right part of Fig. 4) the range of R_s becomes:

$$(28) \quad 22.7\Omega \leq R_s \leq 84\Omega.$$

We see that our fit is wrong by at least a factor of 4. It is clear that some more measurements are needed with a simpler shape of the vacuum chamber to have a chance to find a better

²As the figure has a double logarithmic scale any linear increase will result in a slope of one.

Figure 12: Influence of Vacuum Chamber on Impedance

agreement with a theoretical model. At the same time we will attempt to study the problem with numerical tools.

We will now show the actual measurement of the induced voltage in the magnet. Fig. 13 summarizes different cases with and without the vacuum chamber. For the case without the vacuum chamber we did two measurements: firstly (black solid line in Fig. 13) with a pre-resistor of $R_p = 1000\Omega$ and an input voltage of 1V and secondly (light solid line in Fig. 13) with a pre-resistor of $R_p = 10\Omega$ and an input voltage of 10V. As expected the two curves differ by 20dB and in the second case (see eq. 26) the constant value of U_I is reached at 1000Hz. We therefore choose the first case in which the linear increase of U_I is undisturbed up to about 2000Hz. Introducing the vacuum chamber (black broken line in Fig. 13) a clear deviation is visible already at 200Hz (the same conditions have been applied as in the case without vacuum chamber). From the difference of the two curves we can derive the shielding factor which is depicted in Fig. 14 (black solid line). This can be compared to the damping in the model used to fit the impedance as a function of frequency (broken light line): the agreement is excellent, showing the relevance of the model. We also show that there is practically no damping when the vacuum chamber is not considered in the fitting model (black broken line in Fig. 14)

Figure 13: Measured Induced Voltage

Figure 14: Shielding Factor

3.4 Dipole Chain of the SPS

We are now ready to describe the transmission line measurement of the string of dipole magnets in the SPS.

Figure 15: Powering of the SPS Dipole Chain

Fig. 15 shows how the line is powered: there are twelve stations around the ring which are in series so that each station serves the complete line. We therefore have to measure the transmission characteristics of the total chain of dipoles. For a comparison we have also done the measurement taking only one third of the chain of magnets.

In Fig. 16 the theoretical model is depicted. We are studying the dipole chain with and without the vacuum chamber, the values of the electric components can be found in Tab. 1 (case 4, 5, 6) of the Appendix.

Figure 16: Model of Chain of Dipole Magnets in the SPS

In Fig. 17 the comparison is shown between our model (broken lines) and the transmission line measurements (solid lines) of the full dipole chain with open and closed end (distinguished by light or dark shadings). In this case (appendix, entry 4 in Tab. 1) we have not included the vacuum chamber ($k = 1$ & $R_s = 0$). Over the full range from 10Hz and 1000Hz the agreement is excellent including the location of the resonance frequencies and the prediction of the frequency at which both curves join.

In this fitting the required damping resistor $R_d = 10\Omega$ for the fit is much smaller than the resistors which have actually been installed at each magnet (in fact on average R_d is 25Ω). Here our modelling of the vacuum chamber comes into play: when the R_d is set to 25Ω and we use a shielding resistor of $R_s = 16.7\Omega$ keeping k at 1 (appendix, entry 5 in Tab. 1), the total shielding stays at 10Ω and of course we obtain the same results. Notice that this value of R_s is much closer to the range given in eq. 28. It has to be mentioned that in the real machine the ratio k is likely to be closer to 1 because the vacuum chamber is pressed to fit the pole-faces. In our shielding measurement, however, there had to be free space between vacuum chamber and pole-faces (see right part of Fig. 4) to allow an easy installation and removal of the vacuum chamber which resulted in a k smaller than 1. Moreover there are two types of dipole magnets (MBA and MBB) in the SPS with different free apertures and consequently different shapes of the vacuum chambers. It is therefore no surprise that fit values for the shielding R_s differ for these two measurements.

Figure 17: Amplitude of Impedance for the full Dipole Chain

Figure 18: Amplitude of Impedance with Vacuum Chamber

We have also tried to use a splitting of the inductor by using $k = 0.5$ like in the MBB-s study (appendix, entry 6 in Tab. 1). Fig. 18 shows that in an intermediate regime up to about 200Hz the agreement between model and experiment is even better but at the expense of a deviation at larger frequencies which sets in at about 400Hz. We conclude that in the low frequency regime which is interesting for long transmission lines the splitting plays little or no role. However, the shielding resistor is indeed essential to understand the measurements. We hope that a more thorough study on the vacuum chamber shielding will further clarify these issues.

For comparison we have also studied one third of the SPS dipole chain. In Fig. 19 we present the results of the closed circuit measurement (here expressed in real and imaginary part). As the resonance frequency f_r decreases linearly as a function of the length of the transmission line (see eq. 16) we expect a three times larger frequency. Comparing Fig. 17 and Fig. 19 we find indeed this difference in the resonance frequency f_r , 23Hz and 70Hz for the total chain or one third of the chain respectively.

The most important property of the transmission line concerning tune ripple is the damping as a function of frequency. This damping we express in the number of elements after which the signal is reduced to $\frac{1}{e}$ of the input values. That is: setting l (the length of one element) to 1 we plot $\frac{1}{\alpha}$, where α is the damping of one element of the transmission line. In Fig. 20 we depict the so defined $\frac{1}{\alpha}$, solid lines corresponding to measurements and broken lines to our model. The agreement for the full line of dipole magnets is almost perfect which could have been expected from the good fits of the impedance shown in Fig. 17. The results from one third of the machine (black solid line in Fig. 20) which should in theory be identical, deviate somewhat from the results of the full machine. This probably means that the transmission line is not being built up of completely identical elements. Nevertheless the agreement is quite good. The essential number is the damping frequency f_d at which the input signal at the start of the transmission line is damped to $\frac{1}{e}$ at its end. In our case we have 744 dipoles and we get a damping frequency f_d of 75Hz and 90Hz for the full chain and for one third of the SPS respectively. A summary of all transmission line results is given in Tab. 2 of the Appendix.

Figure 19: Impedance of one third of SPS Dipole Chain

Figure 20: Damping of Ripple Amplitude for the SPS Dipole Chains

3.5 Quadrupole Chain of the SPS

The powering of the quadrupoles is depicted in Fig. 21. The Q_d string (108 units) is powered with one power supply, while the Q_f chain is divided in two equal sections (54 units each). These two are powered in a bridge-like fashion so that by varying the ratio of the two voltages at the power supply each section can be powered differently.

Figure 21: Powering of SPS Quadrupole Chains

This has allowed the so-called tunesplit operation of the SPS. In our case the two sections were powered with the same voltage so that the combined Q_f string can also be considered as one transmission line. It is therefore sufficient to measure only one of the strings of the quadrupoles of which we have chosen the Q_d chain.

For the quadrupoles we have not yet measured the shielding of the field due to the vacuum chamber. However, the results of the study for the dipoles should be applicable here as well.

We expect a small shielding resistor R_s , but the splitting k of the inductor will probably not be important at the low frequencies that are relevant for such long transmission lines. As for the dipoles we incorporate the shielding resistor R_s by lowering the damping resistor R_d . The simple fitting model is shown in Fig. 22 with the parameters of entry 7 in Tab. 1 of the Appendix.

Figure 22: Model of Chain of Quadrupole Magnets in the SPS

For the quadrupoles we have done the same impedance measurements with open and closed circuits as for the dipoles. A comparison of the amplitude between model (dashed lines) and experiment (solid lines) for closed and open circuits are shown in Fig. 23. Besides a different slope at frequencies beyond 2000Hz the agreement is almost perfect. The resonance frequency f_r is 113Hz and about five times larger than the 23Hz for the dipoles.

In Fig. 24 the damping $\frac{1}{\alpha}$ (as in Fig. 20 α is the damping of a unit length element of the transmission line) is shown in units of [magnets] as a function of frequency. The results from the experiment (black solid line) compare well with the model (broken light line) above 100Hz. Below 100Hz the two curves deviate slightly. This is due to the fact that α does not only depend on the amplitude but also on the phase of the impedance which, at low frequency, can not be measured with good precision due the limited sensitivity of our instrument. This problem is not apparent in the case of the dipoles (above section 3.4) because of their five times smaller resonance frequency f_r .

Fortunately the results are fine close to the damping resonance f_d that indicates a damping to $\frac{1}{e}$ of the signal at the end of the transmission line. This damping frequency for 108 quadrupoles is calculated to be 250Hz and 241Hz for the experiment and model respectively. All relevant results concerning the transmission line measurements including the case of the dipoles are shown in Tab. 2 in the Appendix.

Figure 23: Amplitude of Impedance for the Quadrupole Chain

Figure 24: Damping of Ripple Amplitude for the SPS Quadrupole Chain

4 Power Supply and Tune Ripple

Having the transmission line characteristics of the different transmission lines we can now address the task of how to calculate the tune ripple from a measured voltage ripple. An example of the voltage measurement of Q_{f2} is given in the upper part of Fig. 25. It shows the raw signal which corresponds to 500 ± 0.8 mV. (This value has to be multiplied by about 800 to get the actual value in Volts at the power supply.) The lower part of the figure depicts the FFT of the data (0dB corresponds to 1mV). As expected we find a rich spectrum of multiples of 50Hz, the 50Hz and the 600Hz being particularly large and some like 500Hz or 1000Hz hidden in the noise.

The first step is to convert this voltage spectrum into a current spectrum via the impedance of the closed circuit. We have done a current measurement so that we can compare the quality of this first conversion. Fig. 26 shows the comparison for the Q_f case (relative current values are shown in units of [ppm]): above 200Hz the agreement is rather satisfying but there are large differences for the lower frequencies. Of course we would like to avoid one step of the conversion and use the current measurement directly. However, we believe that the differences at low frequency are due to problems in the current measurement itself because we find additional lines of 70Hz and multiples of it. These lines can only come from the current measurement system as they are not present in the voltage measurement. We therefore decided to disregard the current measurement system especially as our first conversion seems to work rather well.

The peaks in the FFT spectrum are used to determine the amplitude of the ripple lines. We have therefore made a careful calibration of our FFT routine and found that between 500Hz and 1000Hz the necessary correction increased from 1% to about 6%.

The next thing to consider is the damping due to the length of the transmission line. It is however not sufficient to simply use the α curve as depicted in Fig. 20, 24, instead what has to be calculated is the current that actually flows on average in the magnet, because only that current can produce a ripple field. We therefore averaged for every multiple of 50Hz the currents of all inductors around the machine thereby taking into account the time dependence of this average.

As the damping is by far stronger for the dipole chain than for the quadrupoles (which has already become apparent in the previous chapters) and as the voltage ripple is not significantly larger in the case of the dipoles, the major contribution to tune ripple comes from the quadrupoles. This also eliminates the effect of partial cancellation of certain ripple lines that would be possible between the quadrupole and dipole chain due to their different resonance frequencies.

Figure 25: Voltage Measurement of Q_{f2} Quadrupole Chain

Figure 26: Relative Current Ripple

Figure 27: Comparison of Horizontal Tune Spectra

Figure 28: Blow-Up of Horizontal Tune Spectrum

Figure 29: Comparison of Schottky Tune Ripple Measurements

Finally we have to consider the conversion from relative current ripple to tune ripple for which we need the natural chromaticity ($\varepsilon = Q' = -33.6$). This is obvious for the quadrupoles but holds true also for the dipoles as long as the chromaticity is corrected by sextupoles and the RF is switched off. These have been the conditions in which we have measured the tune ripple with the Schottky system. In Fig. 27, 28 we compare the tune spectra from the Schottky measurement and from the converted voltage measurement from 1993. With the first graph we want to point out how the tune spectrum would look (black curve with triangular markers) if there would be no transmission line damping (e.g. a huge 600Hz line). Moreover it becomes apparent that around the resonance frequency (113Hz) of the quadrupoles the undamped curve is already very close to the measured tune values because there the damping is small. This is another indication that the ripple is mainly dominated by the quadrupoles. Fig. 28 is a blow-up of the former graph and shows a good overall agreement between the two tune spectra. The large 200Hz, 500Hz and 1000Hz ripple lines in the Schottky spectrum seem not to be caused by power supply ripple but for the time being we do not have any explanation where they might come from.

In the final graph (Fig. 29) we show a comparison of two Schottky tune measurements from 1992 and 1993. The curve of 1992 looks much more like the curve we expect from our voltage measurement, the 200Hz line is smaller by a factor of 10 and the 500Hz & 1000Hz lines are also much less pronounced.

In the experiments of 1994 we will try to clarify the origin of those lines by simultaneously measuring voltage and tune ripple and thereby checking for correlations.

5 Conclusions

The measurements of transmission lines characteristics concerning the SPS dipole and quadrupole chains have been completed and found to be well represented by a simple equivalent electric model. An electric representation of the shielding due to vacuum chambers can explain the very small damping resistor R_d suggested by the fitting. We have therefore concluded that this shielding is sufficiently well considered in our transmission line model. However, more measurements with less complex shapes of vacuum chambers are needed and agreement sought for with simple theoretical models. Moreover, it will be attempted to calculate numerically the shielding effect of the actual vacuum chamber shape in the MBB-s which can be directly compared with our measurements.

The derived tune spectra from the SPS voltage ripple measurements show an overall satisfactory agreement with the tune spectra measured with beam using our Schottky system. There are however frequencies (i.e. 500Hz, 1000Hz) which seem not to originate in the power supplies. Moreover it seems that the spectra vary considerably from measurement to measurement. This year we will therefore attempt to simultaneously measure the tune with beam and the voltage ripple so as to find correlations between the two.

As the whole study is in view of specifying the main power supplies of the LHC we will define a realistic transmission line model for the LHC magnet chains (of course including shielding due to the vacuum chamber). This will lead us to an approximate tune ripple spectrum which we will use in tracking studies to evaluate the combined effect of non-linearities and tune ripple.

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A Appendix

Electric Components of Transmission Lines									
Model	Case #	R_p [Ω]	RL [m Ω]	RC [M Ω]	R_d [Ω]	R_s [Ω]	C [nF]	L [mH]	k
Electric Equivalent	1	10	592.5	150	(10)	–	11	8.896	1.0
MBB-s	2	1.3	3.51	148.8	650	–	8.65	4.6	1.0
	3	1.3	3.51	148.8	500	6.0	25.0	5.57	0.5
Dipole Chain	4	1.3	4.37	148.8	10.0	–	13.4	8.87	1.0
	5	1.3	4.37	148.8	25.0	16.7	13.4	8.87	1.0
	6	1.3	4.37	148.8	25.0	3.8	13.4	8.87	0.5
Quadrupole Chain	7	1.1	14.2	86.4	24.0	–	14.7	13.9	1.0

Table 1: List of Electric Components for various Transmission Lines

For a better comparison we show in each case one unit of the magnet chain. The dipole chain is built up of 744 units and 108 units represent the SPS quadrupole chain. Data are derived from best fits, values in brackets may or may not be considered.

Transmission Line Measurements				
Experiment				Model
Type	#	f_r [Hz]	f_d [Hz]	f_d [Hz]
Dipole	1	16200	–	–
	248	70	90	–
	744	23	75	75
Quadrupole	108	113	250	241

Table 2: List of Results from the Transmission Line Studies

The resonance frequency is denoted by f_r and at the damping frequency f_d the signal at the end of the transmission line is reduced to $\frac{1}{e}$.