A SMALL VERY WIDE BAND ELECTROMAGNETIC PICK-UP STATION

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SUMMARY

In this report, work on a very wide band pick-up station, developed in view of its possible use in the Proton Synchrotron Booster (PSB) is described.

The bandwidth of the electrostatic pick-up units used with cathode followers is in general limited to less than 200 MHz \(^1\). The electromagnetic pick-up unit described here has a higher cut-off frequency: \(f_h \geq 400\) MHz. During the tests, the PSB vacuum chamber was simulated by a cylindrical pipe with circular cross-section; the proton beam was simulated by an axial wire to which pulses were applied, with a rise-time \(\tau_r \leq 0.25\) nsec. All tests have been conducted so far in the laboratory, where the pulse parameters are well under control of the experimenter.

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1. **INTRODUCTION**

The flow of protons in the vacuum chamber of an accelerator induces a current in the walls of the chamber, equal and opposite to that represented by the flow of the protons.

We can think of the vacuum chamber and the (bunched) beam as a coaxial cable of infinite length, carrying short current pulses. This can be simulated by a metal cylindrical pipe with an axial conductor. If the conductor is in the centre of the pipe, the wall current will have a uniform distribution along the circumference. The wall current can be used to produce a signal for observation, through a suitable coupling device; this device is the first element determining the frequency response of the observation system. One of the various possible devices is a wide band transformer, the primary of which will carry the wall current, or a fraction thereof. This can be obtained by cutting a gap in the pipe and connecting a load (a resistor, the primary of a transformer) across the gap, as shown in Fig. 1.

![Fig. 1](image)

If we can be sure that all the current flows through the load, the impedance of the lead would determine both the signal amplitude and the frequency response. In the real machine, however, there are inevitably current paths shunting the gap; as these are not controllable, and practically impossible to analyse fully, their influence can be reduced by making the load impedance as small as possible, compatible with the requirement of having a signal large enough to be easily observed.

For this reason, a transformer is preferable to a resistor as it can be adapted to the characteristics of the transmission line. A single
transformer as shown in Fig. 1 alters the cylindrical symmetry and therefore forces the distribution of current to be distorted; this would cause a distortion of the output signals, in a manner dependent on the position of the beam.

This difficulty can be attenuated by connecting several transformers around the circumference of the pipe (Fig. 2) -- as many as possible in practice. For wide band response, the number of turns in the transformers should be as low as possible; this leads to the arrangement schematically indicated in Fig. 2, where the primary winding is simply a straight wire through the transformer's core.

![Fig. 2](image)

2. EXPERIMENTAL SET-UP

The vacuum chamber was simulated by a brass pipe, cut in two equal length parts; the beam was simulated by an axial wire. The dimensions of the system were the following:

- Inner diameter of chamber \( D_i = 96 \text{ mm} \)
- Outer diameter of chamber \( D_o = 100 \text{ mm} \)
- Diameter of the wire \( D_w = 2 \text{ mm} \)
- Total length of the chamber \( l = 2 \text{ m} \)

The resulting characteristic impedance is \( Z_{\text{op}} \approx 270 \Omega \). As the available generator had an internal impedance, \( R_g = 50 \Omega \), the arrangement shown in Fig. 3 was used. The values of the resistors were:

- \( R_1 = 12 \Omega \)
- \( R_2 = 250 \Omega \)
- \( R_3 = 42 \Omega \).
$R_3$ was made up with six 240 $\Omega$ resistors in parallel, connected symmetrically from the central node ($R_1, R_2, R_3$) to evenly spaced points on the circumference of the pipe. The terminating resistor $R_{cp}$ was built in the same way. The maximum (loaded) output voltage from the generator being 50 V, the maximum current in the pipe was $I_p \approx 75$ mA.

3. THE TRANSFORMERS

The design equations for a wide band transformer are given in Appendix A. The calculation of the upper and lower cut-off frequencies is given in Appendix B.

If we suppose the total load across the gap to be 0.1 $\Omega$, obtained with $N = 10$ transformers, we can write, for one transformer (Fig. A.1),

$$R_b = N \cdot R_{gap} = 1 \Omega.$$ 

If we neglect all losses at low frequencies ($R_p \gg R'_b; \ R'_1, \ R'_2 \ll R'_b$), we can calculate the required inductance for a given lower cut-off frequency. If we use Eq. (A.5), specifying

$$A_1 = 3 \text{ dB at } \omega = \omega_L$$

we find
\[
L_p = \frac{R_b'}{\omega_L} = \frac{R_b'}{2\pi f_L}.
\]

For \( R_b' = 1 \Omega \), \( f_L = 190 \text{ kHz} \) (Appendix B), we have \( L_p = 0.83 \mu \text{H} \). To obtain such an inductance with only one primary turn and small physical dimensions, a material with a very high \( \mu \) is required for the magnetic core. We used a core of Ferroxcube 3B3 (\( \mu \geq 10^4 \)), with the following dimensions:

\[
\varphi_0 = 6 \text{ mm}, \quad \varphi_1 = 4 \text{ mm}, \quad h = 2 \text{ mm}.
\]

This gives

\[
L_p = \frac{4\pi \mu \times 10^{-9} \times (\varphi_0 - \varphi_1) \times h}{\pi (\varphi_0 + \varphi_1)} > 1.6 \times 10^{-6} \text{ H}.
\]

The secondary winding is loaded by a coaxial cable, matched at the far end, with characteristic impedance \( Z_0 = 75 \Omega \). This load, transferred to the primary side, becomes \( R_b' = R_1/n^2 \) where \( n \) is the secondary to primary turns ratio.

To have \( R_b' = 1 \Omega \) we used \( n = \sqrt{75}/1 \approx 8 \). This is also the number of secondary turns. The secondary inductance will be \( L_s = 64 \times L_p \geq 103 \mu \text{H} \). The measured value was \( L_s = 110 \mu \text{H} \), which is in agreement with the calculation. This value is twice as large as the required value. The actual cut-off frequency will be, therefore:

\[
f_L = \frac{1}{2\pi \times 1.6 \times 10^{-6}} \approx 100 \text{ kHz}.
\]

Wire of rather large diameter was used (2 mm for the primary, 0.3 mm for the secondary) to keep resistances negligible.

The secondary winding was distributed as uniformly as possible on the circumference of the toroid.

Calculation of the upper cut-off frequency was not attempted because of the difficulty of knowing the value of the capacitances involved, and the leakage inductances. Care was taken to have short leads, even spacing of turns and uniformity between transformers; the determination of the frequency response was made experimentally.
Fig. 4

A split collar with 10 BNC connector's holes
Fig. 5

Second split collar without holes
4. CONSTRUCTION OF THE PICK-UP UNIT MODEL

The mechanical assembly consisted of two split collars, one of them bearing 10 BNC connectors (see Figs. 4 and 5). The collars were accurately fitted to the outside diameter of the pipe, so as to ensure good contact when tightened around it. Each collar was appropriately indented for easy positioning of the rigid wires constituting the primary windings of the transformer.

The assembling was done in two halves, each complete with its five transformers. The primary wires were soldered in their place, the secondary soldered to the connectors, and finally the two halves were joined around the pipe to form the complete unit.

5. EXPERIMENTAL RESULTS

The signals from the transformers were fed, through 75 Ω cables, to a passive summing circuit (Fig. 6).

![Diagram of summing circuit](image)

The summing circuit provides a good match between each cable and the common load, ensuring at the same time insulation between any two transformers.

If we consider as a reference the case where each transformer would be terminated on 75 Ω, and only one is used to get an output signal, the summing circuit introduces a loss of about 4.7 dB. Its usefulness consists in the fact that the output signal is substantially insensitive to the beam (wire) position relative to the centre of the vacuum chamber (pipe).
Fig. 7a
Pulse from generator
Vertical 50 mV/cm: Horizontal 5 nsec/cm

Fig. 7b
Leading edge of pulse from generator
Vertical 50 mV/cm: Horizontal 0.5 nsec/cm
**Fig. 8a**

Pulse from transformer  
Vertical 20 mV/cm: Horizontal 5 nsec/cm  
Generator output $V_g = 50$ V

**Fig. 8b**

Pulse from transformer (leading edge)  
Vertical 20 mV/cm: Horizontal 0.5 nsec/cm  
Generator output $V_g = 50$ V
Fig. 9a

Output from summing circuit
Vertical 10 mV/cm: Horizontal 5 nsec/cm
Generator output $V_g = 45 V$

Fig. 9b

Output from summing circuit (leading edge)
Vertical 10 mV/cm: Horizontal 0.5 nsec/cm
Generator output $V_g = 45 V$
The generator used was a Tektronix, type 109, which gives variable length pulses, with a rise-time $T_r \leq 0.25$ nsec, and maximum amplitude of 50 V.

To display the pulses a sampling oscilloscope (Tektronix type 561) was used. Figures 7a and 7b show the pulse from the generator and its leading edge, respectively.

Figures 8a and 8b show the pulse at the secondary of one of the 10 transformers, and its leading edge, respectively. The first (positive) peak seen on the top of the pulse in Fig. 8a, and the negative step immediately following, are due to imperfect match at the end of the pipe. A possible way of improving the matching would be the use of a disc resistor rather than the star arrangement of six resistors that we used. The second peak (negative) is due to the pulse generator, and can be observed also on Fig. 7a.

The third (negative) peak is also due to mismatch at the end of the pipe. The series of peaks after the pulse is an image of those occurring on the top. Figures 9a and 9b show the output pulse from the summing circuit and its leading edge, respectively. By comparison with Fig. 8 one can see that the degradation is rather small. The rise-time is larger than that of the pulse generator, but shorter than 1 nsec, indicating an upper cut-off frequency of the system $f_h > 350$ MHz. Figures 10a and 10b show the leading edge of the same pulse observed after transmission through two cables, of lengths $t_1 = 1$ m and $t_2 = 21$ m. The longer cable was a high quality 75 Ω cable. It can be seen that even in this case the rise-time is shorter than 1 nsec. The attenuation of the pulse is less than 15%.

The cut-off frequency of the system is related to the time constant $T_L$. $T_L$ can be measured by observing the decay of comparatively long pulses. To do this we used a pulse generator, model EH 139 B. A pulse of length $T_P \approx 7$ µsec (see Fig. 11) is differentiated. The time constant is found to be $T_L \approx 2$ µsec, corresponding to $f_L \approx 80$ kHz.

6. INFLUENCE OF A CONDUCTING SHIELD

The above tests were performed with the pick-up station as described. However, in the real machine, the station will be enclosed in a conducting shield, to protect it from electrical interference, and to provide a controlled current path shunting the pick-up gap. Both functions are rather
Fig. 10

Leading edge of output pulse after transmission through a 75 Ω cable

a) Cable length $t_1 = 1$ m; b) Cable length $t_2 = 21$ m

Vertical 20 mV/cm; Horizontal 0.5 nsec/cm

Fig. 11

Pulse from summing circuit: low frequency response without shielding box.

Vertical 5 mV/cm; Horizontal 1 μsec/cm. $\tau_L \approx 2$ μsec.
unimportant in the laboratory model, but essential in the real device, where the interference level is very high, and there are many unpredictable current paths across the gap. (In addition, the gap in the real device must be bridged by an insulating vacuum tight wall, to preserve vacuum inside the chamber.)

Further tests were performed on the model shielded by a simple metal box: no appreciable effect was observed on the high-frequency response: $f_L$ was still greater than 350 MHz. (Compare Fig. 12, showing the output pulse with the box, with Fig. 9 showing the same pulse without box.) A marked influence, as it should be expected, was observed on the low-frequency response. This is visible in Fig. 12 and seen very clearly on Fig. 13, which shows that the time constant $\tau_L$ is $\approx 0.4 \mu$ sec, corresponding to $f_L \approx 400$ kHz.

![Fig. 12](image)

Output pulse from summing circuit with shielding box.
Vertical 10 mV/cm:
Horizontal a) 0.5 nsec/cm  
b) 10 nsec/cm.
Pulse from summing circuit: low frequency response with shielding box.
Vertical 5 mV/cm;
Horizontal a) 1 µsec/cm 
b) 0.5 µsec/cm

\[ \tau_L \approx 0.4 \mu\text{sec}. \]

We tried to improve the situation by increasing the impedance of the box as seen from the gap: this was obtained by loading the box with four big rings of magnetic material (ferrite toroids with \( \Phi_L = 100 \text{ mm}, \Phi_O = 180 \text{ mm}, h = 12.5 \text{ mm}, \mu \approx 200 \)). The result is shown in Fig. 14: the time constant is approximately the same as in Fig. 11 (pick-up without box, \( \tau_L \approx 2 \mu\text{sec} \)).

The pick-up unit with and without box is shown in Fig. 15 and Fig. 16. Figure 17 is a close-up view of the gap and the transformers.

7. CONCLUSIONS

From the laboratory tests one can conclude that the pick-up unit described could be used for observation of the shape of the proton bunches in the PSB. Nevertheless, the signal amplitude is rather small, and the signal in the PSB will not be repetitive.
Fig. 14

Pulse from summing circuit: low frequency response with shielding box loaded with ferrite rings. Vertical 5 mV/cm: Horizontal a) 0.5 µsec/cm b) 1 µsec/cm

Fig. 15

The pick-up unit with its shielding box
Fig. 16
The pick-up unit without shielding box

Fig. 17
Close-up view of the pick-up unit
The dynamic range of the beam current in the PSB will be from about 30 mA to 2 A per ring. This is the peak to peak value of the fundamental component, the minimum being for a low-intensity beam at the beginning of acceleration, the maximum for a high intensity beam \((2.5 \times 10^{12}\) protons per ring) at the end of acceleration. With the device described this corresponds to an output signal from about 13 mV to 0.9 V.

Since the signal is not repetitive, a sampling oscilloscope cannot be used. Existing wide band scopes with real time display (HP 183 A, Tektronix 619) allow photographic observation of single pulses, but direct viewing at high speed is problematic.

As far as the sensitivity is concerned, the HP 183 A is adequate for all but the smallest signal (vertical sensitivity 10 mV/cm) if the limitation in the bandwidth is acceptable (\(\approx 250\) MHz). The bandwidth can be increased (beyond 500 MHz) if one drives the CRT deflection electrodes directly: to do so, however, would require a very wide band external amplifier (vertical sensitivity 3 V/cm).

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The pulse transformers required for the electromagnetic pick-up unit must have a very wide band response. The equivalent diagram of such a transformer is given in Fig. A.1. All secondary components have been transferred to the primary side.

\[ \text{Fig. A.1} \]

- \( E_a \) is the source e.m.f.
- \( R_a \) is the source resistance.
- \( R_1 \) is the resistance of the primary winding.
- \( R'_s \) is the resistance of the secondary winding referred to the primary side.
- \( R'_b \) is the lead resistance referred to the primary side.
- \( R_p \) is the shunt resistance representing the total loss in the core.
- \( C_1 \) is the effective capacitance shunting the terminals of the primary winding.
- \( C'_s \) is the corresponding secondary capacitance referred to the primary side.
- \( L'_1 \) is the primary leakage inductance.
- \( L'_s \) is the secondary leakage inductance referred to the primary side.
- \( L_p \) is the open circuit inductance of the primary.

We are interested in its study from the point of view of bandwidth and losses.
The relations between the equivalent circuit of a transformer and its transmission and reflection characteristics are a special case of the more general passive network theory.

The insertion loss, \( A_1 \), is the ratio, in dB, of the power in the load without the network to the power in the load when the network is inserted, i.e.

\[
A_1 = 20 \log_{10} \left( \frac{\text{load voltage when network is replaced by direct connection}}{\text{load voltage with network}} \right) [\text{dB}] . \quad (A.1)
\]

The effective loss, \( A_e \), of a network is the ratio in dB of the applied power to the power in the load when the network is inserted, i.e.

\[
A_e = 20 \log_{10} \left( \frac{\text{load voltage when network is replaced by ideal matching transformer}}{\text{load voltage with network}} \right) [\text{dB}] . \quad (A.2)
\]

The effective loss is normally used to describe the attenuation of a transformer since it compares it to an ideal matching transformer. In our case of a current transformer it is preferable, however, to express the behaviour in terms of insertion losses.

The three main regions of the transmission characteristic will be considered. For each region the insertion loss will be quantitatively related to the elements of the equivalent circuit.

1. The mid-band region

The insertion loss due to the series resistance is:

\[
A_1 = 20 \log_{10} \left( 1 + \frac{R_s}{R_a + R'_b} \right) [\text{dB}] , \quad (A.3)
\]

while for the shunt resistance:

\[
A_1 = 20 \log_{10} \left( 1 + \frac{R}{R_p} \right) [\text{dB}] \quad (A.4)
\]
where

\[ R_s = R_1 + R_2 \]

\[ R = R_a R_b / (R_a + R_b) \]

2. The low frequency region

In general, the attenuation in the frequency response at low frequencies is due to the shunt impedance. This impedance decreases as the frequency decreases and causes a progressive increase in attenuation, which adds to the mid-band loss. In most cases the contribution of \( R_p \) to this attenuation is negligible. Then:

\[ A_i = 10 \log_{10} \left[ 1 + \left( \frac{R}{\omega R_p} \right)^2 \right] \text{ [dB]} \]  \hspace{1cm} (A.5)

3. The high frequency region

The response at high frequencies is influenced mainly by the leakage inductance and the shunt capacitances, and their relations to \( R_a \) and \( R_b \). In a low impedance circuit the series leakage reactance may be appreciable, while the shunting effect of the capacitance could be negligible, while the reverse could apply to a high impedance circuit.

The high frequency attenuation due to leakage inductance is given by:

\[ A_i = 10 \log_{10} \left[ 1 + \left( \frac{\omega L}{R_a + R_b} \right)^2 \right] \text{ [dB]} \]  \hspace{1cm} (A.6)

and that due to shunt capacitance is given by:

\[ A_i = 10 \log_{10} \left[ 1 + (\omega CR)^2 \right] \text{ [dB]} \]  \hspace{1cm} (A.7)

where \( C = C_1 + C_2 \).
DETERMINATION OF THE LOWER AND HIGHER CUT-OFF FREQUENCY REQUIREMENTS

At injection the radio-frequency of the PSB is $f_1 = 2.996$ MHz and the bunch length in electrical degrees is $\vartheta = 240^\circ$. The corresponding values at ejection are $f_2 = 8.15$ MHz and $\vartheta = 145^\circ$. The corresponding time duration of the bunches and gaps are:

\[
T_{1p} \approx 220 \text{ nsec}, \quad T_{ig} \approx 110 \text{ nsec at injection}
\]
\[
T_{2p} \approx 50 \text{ nsec,} \quad T_{2g} \approx 75 \text{ nsec at ejection}.
\]

If we suppose, as a crude approximation, that the bunches give rise to rectangular pulses, the situation will be that depicted in Fig. B.1.

\[
\Delta V
\]
\[T_p + T_g = T; \quad v' + v'' = v \]

Fig. B.1

The time constant $r_1$ of the transformer causes the flat portion of the pulse to sag by an amount $\Delta V$. If $r_L$ is long compared with $T$, we can approximate the exponential with a straight line and write:

\[
\frac{\Delta V}{V'} = \frac{T_p}{r_L} \quad \text{and} \quad \frac{V}{V'} = \frac{T - T_p}{T},
\]

from which

\[
\Delta V = V' \cdot \frac{T_p}{r_L} = V \cdot \frac{T_p}{r_L} \left( \frac{T - T_p}{T} \right).
\]
If we denote by $k$ the relative loss $\Delta V/V$, we have:

$$k = \frac{\Delta V}{V} = \frac{T_p (T - T_p)}{r_L T}.$$  \hspace{1cm} (B.1)

The cut-off frequency $f_L = 1/2\pi r_L$ of the transformer can be written as:

$$f_L = \frac{k T}{2\pi T_p (T - T_p)}.$$  \hspace{1cm} (B.2)

For given values of $T$ and $T_p$, $k$ will depend on $f_L$; if we fix an upper limit for $k$, this will give an upper limit for $f_L$.

The worst case, for pulses of variable electrical length, occurs when

$$T_p = \frac{T}{2}.$$  

In that case

$$f'_L = \frac{k}{\pi T_p} = \frac{2k}{\pi T} = \frac{2k}{\pi} f.$$  \hspace{1cm} (B.3)

If we admit a value of $k = 0.1$ in the worst case, we must have

$$f'_L \leq \frac{0.1 \times 2 \times 2.99 \times 10^6}{\pi} \approx 1.92 \times 10^6 = 192 \text{ kHz}.$$  \hspace{1cm} (B.4)

The upper cut-off frequency determines the shortest rise-time that can be obtained: if we want to transmit pulses with a rise-time $r_r \approx 1 \text{ nsec}$, we must have

$$f_h \geq \frac{2.2}{2\pi r_r} = \frac{0.35}{10^{-9}} = 350 \text{ MHz}.$$  \hspace{1cm} (B.5)
The vacuum chamber of the synchrotron was simulated in the laboratory by a copper pipe of internal diameter \( D_i = 96 \, \text{mm} \), and the proton beam by an axial wire of diameter \( D_w = 2 \, \text{mm} \). The characteristic impedance of this coaxial transmission line is:

\[
Z_c = 60 \log_e \frac{D_i}{D_w} = 250 \, \Omega.
\]

The actual value was found by measurement to be 270 \( \Omega \). Once the coaxial line is matched at both ends, the equivalent circuit of the model with the transformers across the gap is that of Fig. C.1.

\[ R_c = 270 \, \Omega. \]  
\[ T_r \] represents \( N \) transformers connected in parallel. \( I \) is the total current flowing in the wall of the pipe. This means that as far as the transformers are concerned, the internal resistance of the generator is \( 2R_c = 540 \, \Omega \).

If there are \( N \) transformers in parallel across the gap, the current in each transformer is \( I_1 = I/N \). The voltage \( V_p \), however, is determined by the drop across \( 2R_c \), carrying the total current \( I \). We have:
\[
\begin{align*}
V_p &= V_a - 2NI R_c \\
I_p &= I_c
\end{align*}
\]

This means that (as long as all loads are symmetrical and no current circulates between transformers) each transformer sees a source impedance \(R_a = 2NR_c = 5500 \, \Omega\).

REFERENCES

