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The new front end module of the TTF strip-line BPM detector with single bunch response.

A new front end of the of the TTF strip-line BPM detector has been built and tested, in order to ensure single bunch response when Injector II will be operating in TTF with 1 MHz bunch repetition rate. The actual beam structure with Injector I consists of long trains (≈ 1 ms) of low charge bunches separated by 4.6 ns. The actual front end has therefore been designed for quasi CW operation and the filters in the RF module have a very narrow bandwidth. The actual detector is made of two separated modules (see Fig.1) : an RF module which selects and converts the 216.7 MHz harmonic of the strip-line signal down to 50 MHz and then amplifies it ; an amplitude to phase (AM-PM) detection module (see Appendix A) which performs the amplitude to phase conversion at 50 MHz and then detects and amplifies the phase signal[1].We recall that the advantage of the AM-PM detector is that the output signal depends only on beam position , i.e. it is independent of beam bunch charge within a wide span and has a wider dynamic range than circuits based on sampling and elaboration of individual electrode voltages [2] .

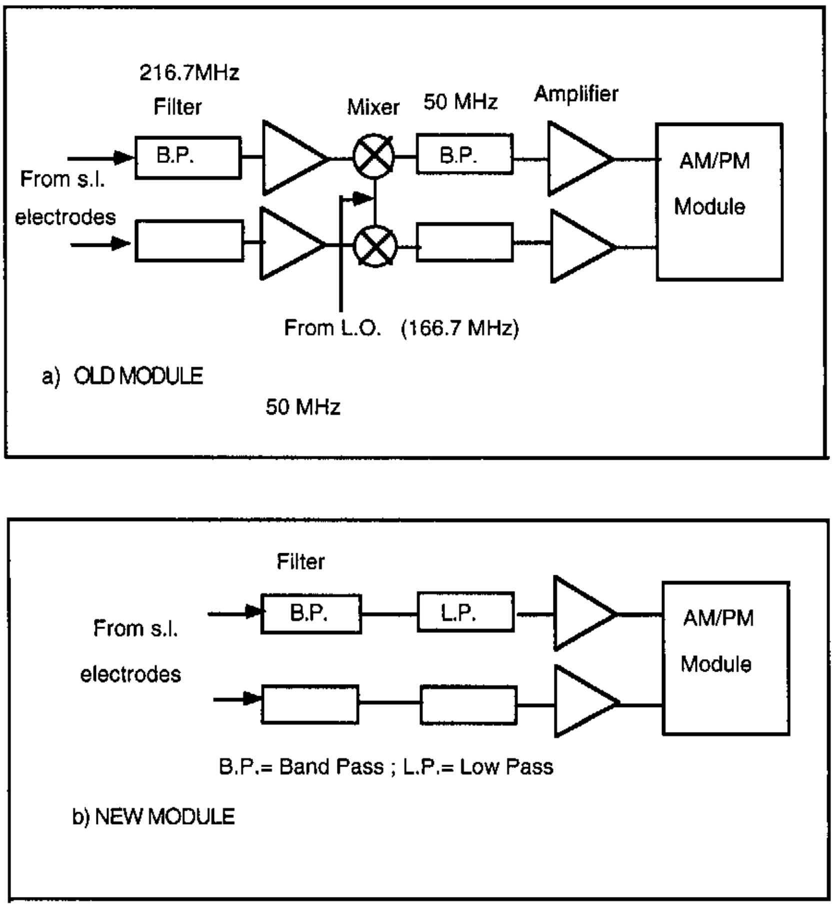

Fig 1- Old and new RF modules.

The beam structure of Injector II has bunches of charge 8 nC and duration less than 10 ps, separated by 1 μs. In order to detect a single bunch the response of the detector should be a fraction of a μs. The bandwidth of the system has therefore to be increased.The bandwidth of the AM-PM module can be increased as necessary with minor component substitutions. The RF module instead needs complete remaking.

It has been verified on the actual RF module that the chains comprising the frequency down-converter do not work well on transients and introduce intolerable fluctuations on the phase detector output. In fact for transient excitation this output is very sensitive to inequalities between the two channels which have to be compared. One should make these channels as equal as possible , therefore they must be simple, with few filters and components that have to be matched.

The simplest solution is the direct excitation of a single band-pass filter, tuned to the final low frequency (in our case 50 MHz), by the short bipolar pulse from the strip- lines.This solution has already been adopted on other accelerators [3],[4].

In Fig.2 are shown the waveforms of the impulse response of filters of the two channels superimposed.

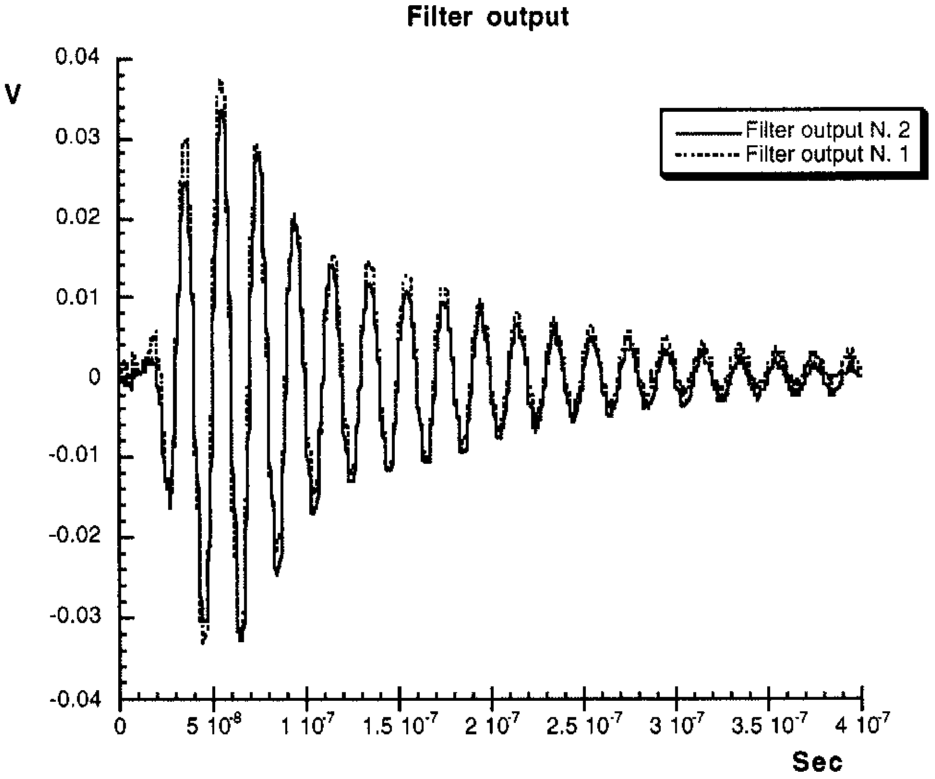

Fig.2- Filter impulse responses.

The peak impulse response of a narrow band filter can be estimated by the following formula, which can be directly justified considering the equivalent low pass Gaussian filter with bandwidth Ω. The impulse response of such a filter has a peak value $\approx \Omega/\pi$ [5]. Therefore:

$$V_p \approx 2\,\Delta f\,Q\,Z_t$$

where Q is the bunch charge , Z_t is the transfer impedance of the strip-lines at 50 MHz, and Δf is the frequency bandwidth.
The pulse duration at half height is about [5]:

$$T \approx 8(\ln 2)^2 / \Omega = 0.6 / \Delta f$$

The transfer impedance is given by:

$$Z_t = k\,Z \sin(\omega L/c)$$

where k is a coupling coefficient . For the simple geometry of a strip occupying an angle Φ of the of the vacuum chamber perimeter $k \approx \Phi/2\pi$. With the complex geometry of the TTF monitor the measured $k \approx 0.03$ [6]. Z is the impedance of the strip-line = 50 Ω, L is the length of the strip-line = 0.17 m. At 50 MHz one gets $Z_t \approx 0.3\ \Omega$, a rather low value; however, with Q=8 nC and $\Delta f \approx 5$ MHz, one still obtains $V_p \approx 25$ mV. Even taking into account the insertion loss of the filter, a moderate amplification by about a factor 10 is sufficient to raise this voltage to the level needed by the successive module. In order to ensure a controlled and constant amplification factor we have adopted an operational amplifier in feedback configuration.We chose the CL425, which is a wideband low noise current feedback amplifier, because its large gain bandwidth product $\approx$1.7 GHz allows to achieve the required gain in a single stage amplifier. Such a large bandwidth slightly deteriorates the noise characteristics of the module, but still within a tolerable level.

The filters must be able to withstand an input voltage of approximately triangular shape with about 70 V peak and about 150 ps duration at 1 MHz repetition rate .In fact the peak voltage directly from the strip-line electrodes is higher ($\approx$ 1 kV) and has a shorter duration (a few ps) but the waveform is stretched and the peak decreased by the $\approx$ 30 meters of cable connecting the BPMs to the detector rack (see Appendix B). Moreover, stray capacitances will further decrease the peak values.
Each couple of filters must have identical center frequencies and coincident roll-off curves and these characteristics must remain constant in time.
The choice of the bandwidth is a compromise between peak voltage and pulse duration requirements, as it is evident from the above formulas.

Commercial filters at such a low frequency and with such characteristics are very bulky and expensive. We have opted for a simple solution consisting of a $\lambda/2$ resonant coaxial cable filter followed by a low-pass filter. The overall filter transfer curves are shown in Fig. 3. They show a good similarity within the 3 dB bandwidth and drifts are minimized by proximity of the cables.

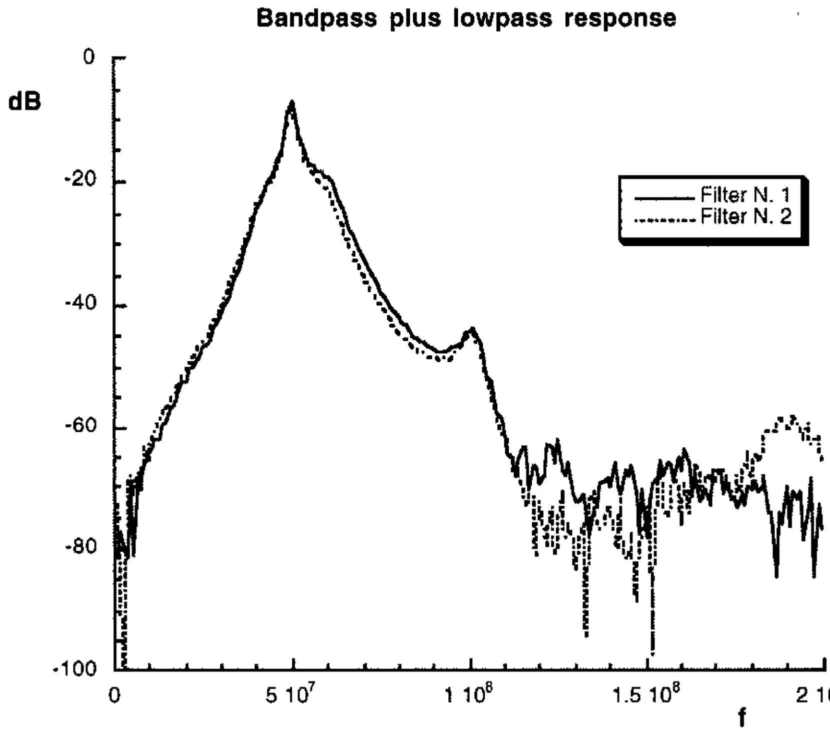

Fig.3- Overall cascaded filters response.

The overall detector output waveform is shown in Fig.4. It is to be remarked that the initial part presents a transient oscillation and the final part has fluctuations due to the

RF filtered signal levels dropping below the limiting threshold, so the sampling of the waveform must be performed within a middle interval spanning about 100 ns .

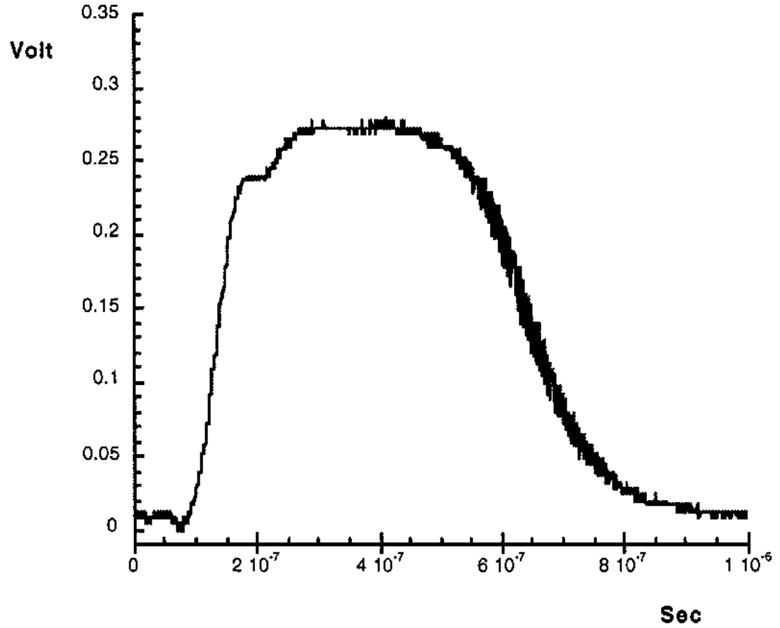

Fig. 4. Output pulse for a 3 dB unbalance of stripline signals corresponding to ≈ 4 nC / bunch @ 1 MHz.

The response of the circuit to a burst of 1 nC bunches at 9 MHz repetition rate, as it is foreseen in the FEL application, is the one shown in Fig. 5. The step time constant is about 200 ns and there is a residual ripple on the steady state , which should cause no troubles because the sampling is made at fixed positions on the waveform.

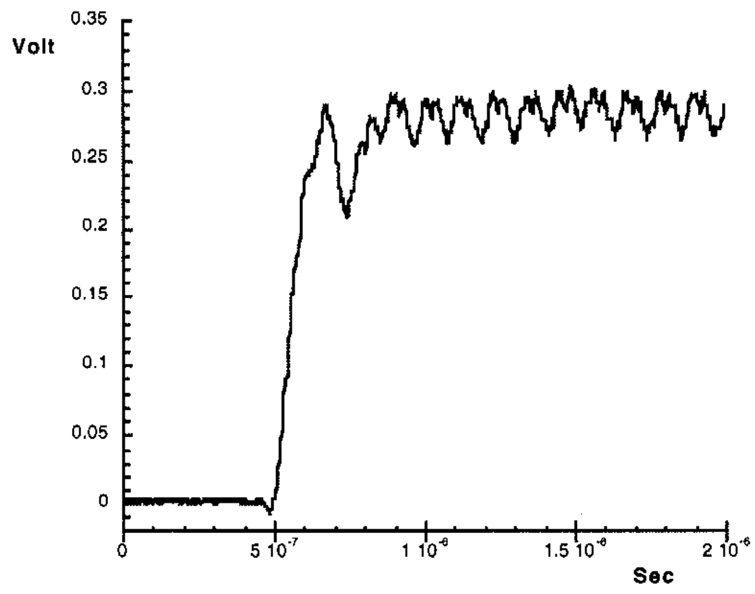

Fig. 5- Response to a 9 MHz burst of 1 nC / bunch.

For small displacements Δx from the center of the vacuum chamber the following relation with the voltage difference between the electrodes holds:

$$\Delta x \approx R/2\,(V_1-V_2)\,/\,(V_1+V_2)$$

where R is half the radius of the monitor vacuum chamber. In our case R=30 mm. The response curve (detector output versus beam displacement) is shown in Fig.6. For large displacements the output curve becomes nonlinear. The waveform of Fig.4 has been sampled at 300 ns from the start. Care must be taken that the voltage level of neither of the two channels falls below the working threshold of the AM/PM module.For balanced inputs (centered beam) this threshold has been measured to correspond to an equivalent beam charge of ≈ 1.2 nC.

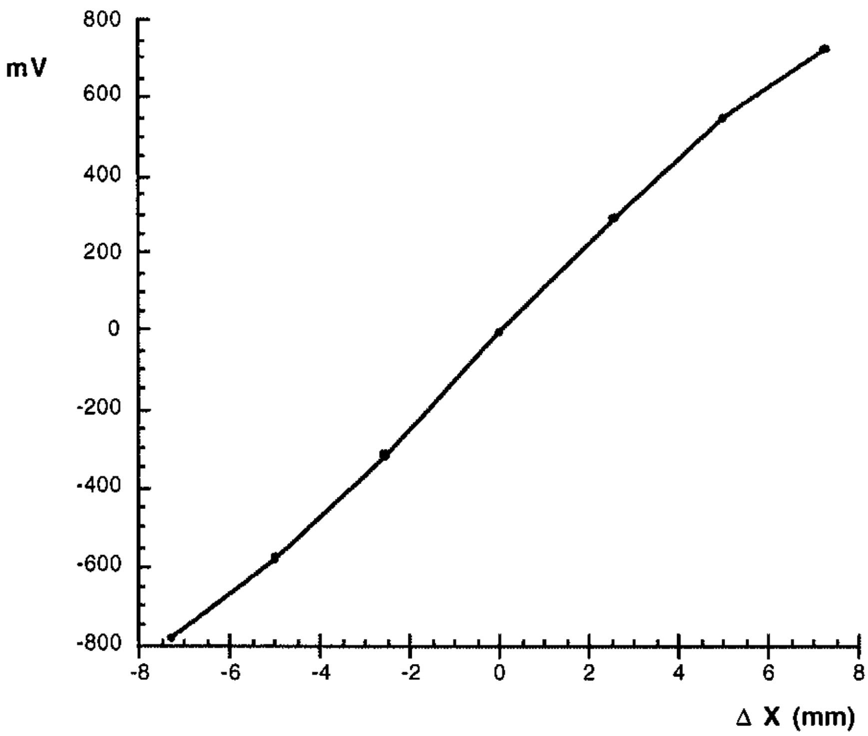

Fig.6. Detector output vs. beam displacement.

In order to evaluate the output noise level we have schematized the sampling circuit as a switch followed by a resistor-capacitor group with a time constant equal to one third of the interval over which the capacitor charges to the new holding voltage (acquisition time).Assuming the acquisition time to be 100 ns, we have computed the final holding voltage of 10 successive detector waveforms (for an input pulse equivalent to 4 nC).
In Fig. 7 is shown a typical noisy waveform, acquired with an oscilloscope, over a time interval of 100 ns on the flat top of the pulse.

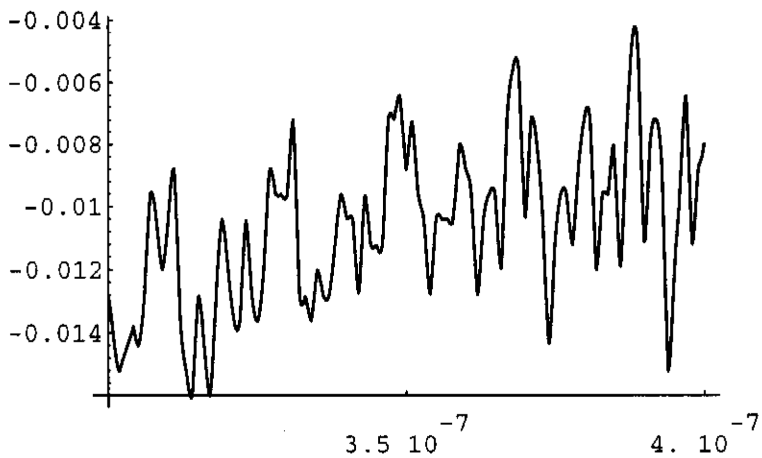

Fig. 7- Typical noisy waveform (V, sec).

In Fig. 8 is shown the result of manipulation of the above waveform by the sampler. The noise is smoothed out by the integrating action of the sampler and the final value is considered as representative of beam position . The sampler input is set to zero before the successive acquisition cycle.

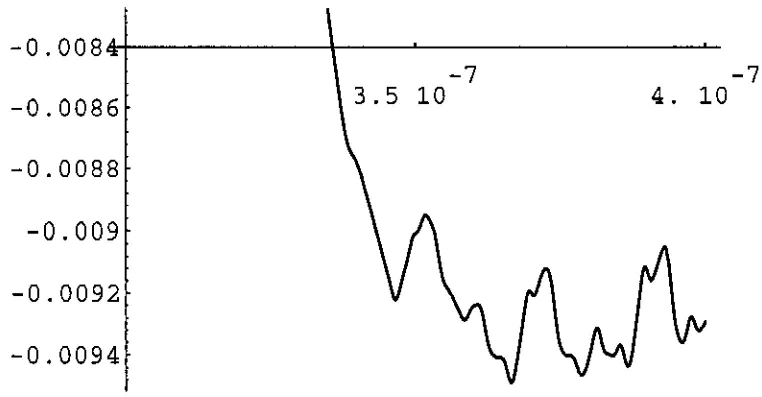

Fig. 8 - Sampler output waveform (V, sec).

The variance of the ten final values is $\approx$ 2 mV, which corresponds to an RMS space resolution of $\approx$ 20 μm.

The lowest detectable beam charge can be decreased below 1 nC by increasing the amplification after the filter, but when switching to 8 nC bunch charge, an attenuator must be inserted ahead of the amplifier to avoid signal distortion. The circuit can be employed both for Injector II and for FEL operations, but attenuators must be inserted or removed from each chain by suitable jumpers or switches, according to the use.

APPENDIX A.

The AM/PM circuit.

We recall briefly the working principles of the amplitude to phase (AM/PM) conversion circuit.

The main blocks of the circuit (Fig.A1) are :
1) a hybrid junction which splits the signals A and B, then delays and recombines them so as to obtain two voltages A+iB and B+iA whose phase difference, for small beam displacements , is approximately proportional to the displacement itself.
2) Two limiting amplifiers which eliminate amplitude dependance of the phase detector ouput and
3) a phase detector.

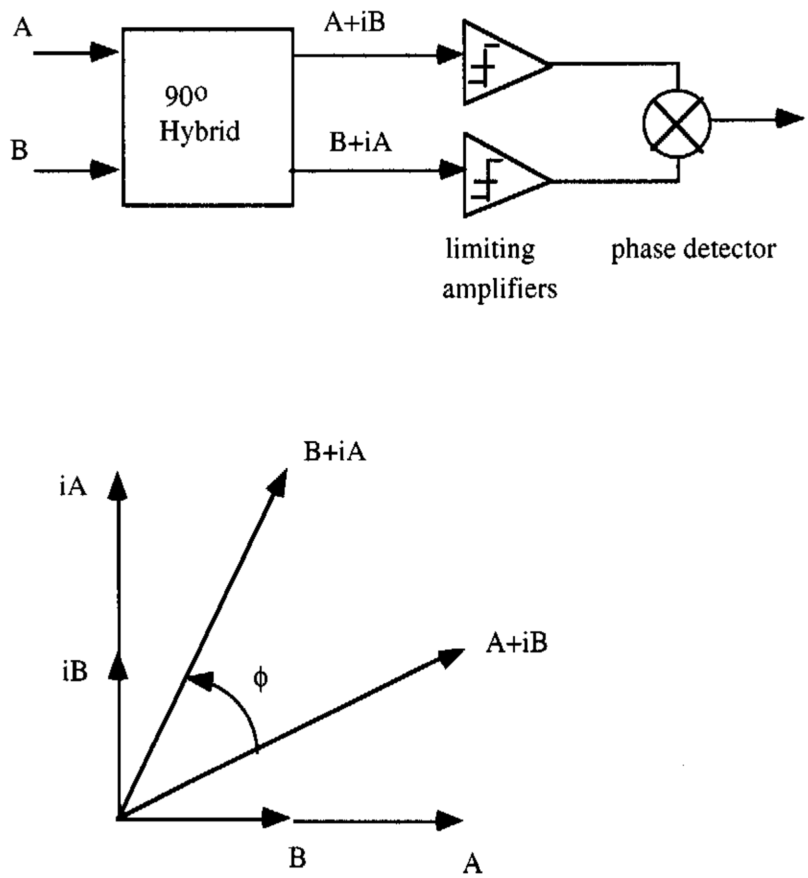

Fig. A1- Scheme of the AM/PM circuit.

With reference to Fig. A1, the 90° hybrid operates as a phase modulator. Its input signals are in direct proportion to the amplitude of voltage signals from a pair of strip lines. The beam position information Δx is related to the signal amplitudes (A,B) by:

$$\Delta x = R(A-B) / 2 (A+B)$$

where R is approximately the radius of the monitor vacuum chamber. From the figure we obtain the relation:

$$(A-B) / (A+B) = \mathrm{Tan} (\phi / 2)$$

so, for small displacements, Δx is proportional to the phase angle ϕ .

APPENDIX B.

The input and output pulse waveforms of a length l of cable connected to the strip-line electrode can be approximately evaluated as follows.

The peak current corresponding to a rectangular bunch of charge Q=8 nC , duration T=10 ps is $I_p = 800$ A.

The peak voltage on the electrode (neglecting stray capacitance) is:

$$V_p = k\, Z_o\, I_p = 1.2 \text{ kV}$$

with k= 0.03, $Z_o = 50\ \Omega$.

The impulse response of coaxial cable of length l, attenuation α (rad/unit length) at frequency f is [7]:

$$g(u) = A\, u^{-3/2}\, e^{-B/u}$$

with u= t-t_d, t_d= delay time of cable, B= π A^2 = $(l\, \alpha)^2 / 4\pi f$

$\alpha = 0.115\ \alpha^*$(dB/unit length)

Assuming a*= 4 dB/ 100 ft @ 400 MHz, l= 30 m , the response (shown in Fig. B1) is:

$$v(u)= k\, Q\, Z_o\, g(u)$$

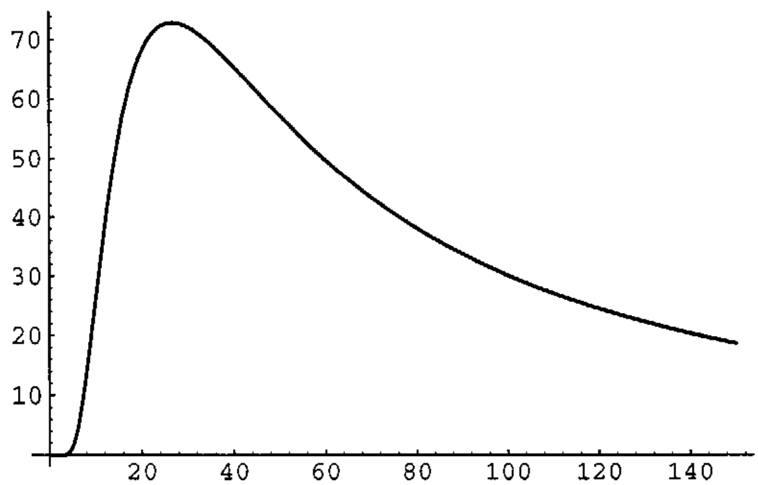

Fig. B1- Cable impulse response-Ordinates in V, abscissae in ps.

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