 IEEE Transactions on Nuclear Science, Vol. NS-32, No. 5, October 1985

BEAM IMPEDANCE MEASUREMENT BY THE WIRE METHOD USING A SYNTHETIC PULSE TECHNIQUE

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Summary

The coaxial wire method is widely used to simulate the image current and the electromagnetic field of a bunch. Measurements are usually performed in frequency (impedance) or time-domain (k-parameter). Some limits of validity and possible sources of error applying this simulation method are discussed. In general one may expect correct results if measurements are restricted to single localized impedances much smaller than the characteristic impedance of the coaxial line (beampipe with center wire). In case of more than one localised small impedance in a given beampipe time-filtering can be applied, provided there is sufficient spatial separation to reduce the problem to the case of isolated single impedances and to eliminate multiple reflections from nonmatched 50 ohm transitions at the end of the beampipe. The technique consists in generating a synthetic pulse in time domain via FFT, from measurements taken in the frequency domain. This leads to higher spectral power density than realtime or sampling pulse measurements thus giving higher dynamic range and better reproducibility. The impedance $Z(\omega)$ as well as the loss parameter k as a function of the bunch length ℓ can then be deduced by computations.

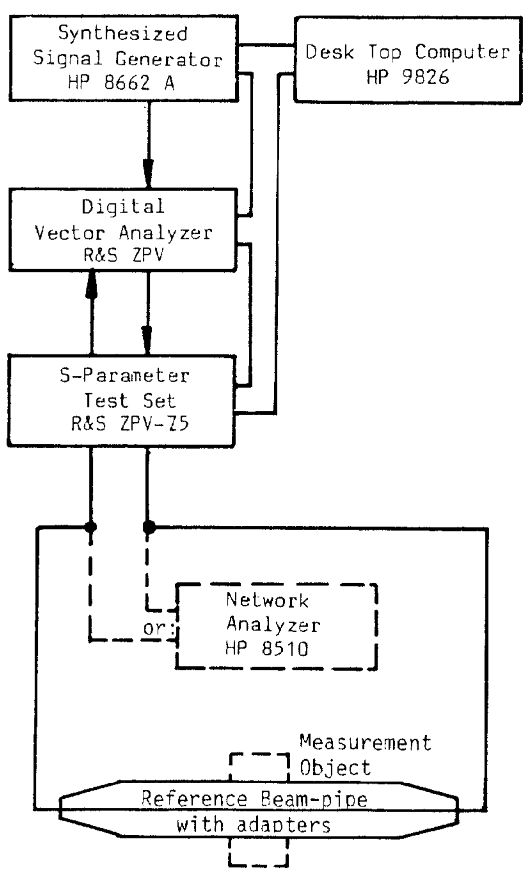

Fig. 1 Measurement Set-up

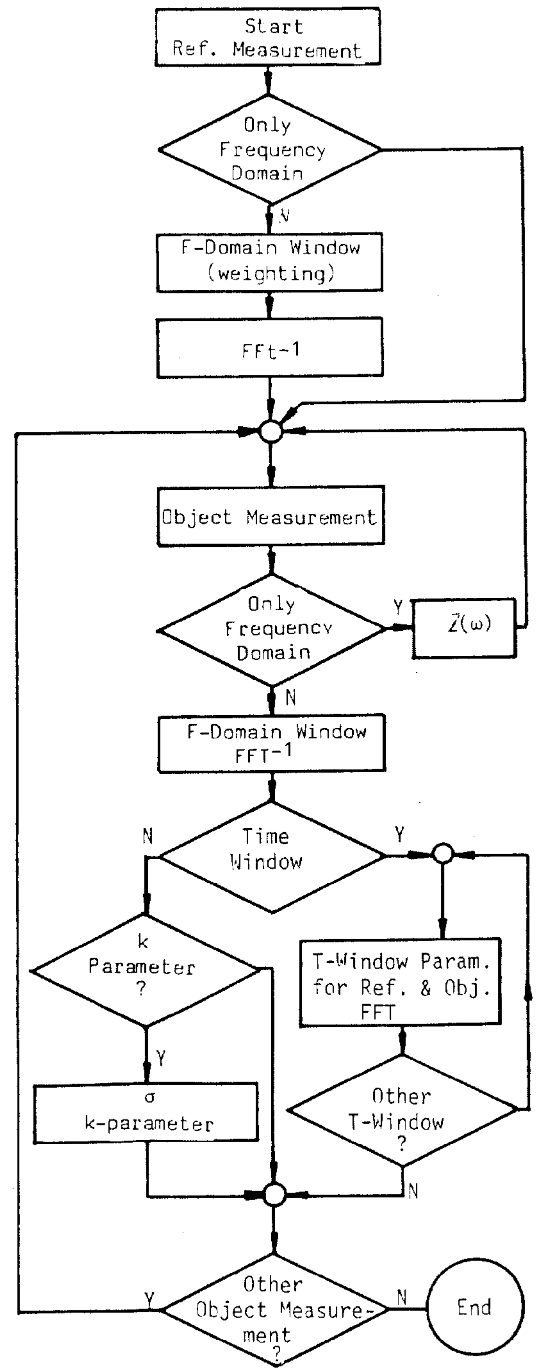

Fig. 2 Simplified Flow Diagram for Measurement Procedure (no k-parameter with HP 8510)

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Measurement Set-up

As depicted in Fig. 1, two independent systems have been used, both based on the classical coaxial wire method [1] which transforms the beam-pipe to be measured in a coaxial transmission line of characteristic impedance Z_L for each considered frequency. The corresponding beam impedance $\underline{Z}(\omega)$ can be related (1) to the scattering matrix transmission coefficient $\underline{S}_{21}(\omega)$. The equivalent two-port is described referring to a reference line with identical matching sections and equal length ($\underline{Z}$=R+jX; $|\underline{Z}|\ll Z_L$).

$$\underline{S}_{21\ OBJ}/\underline{S}_{21\ REF}=(1-R/2Z_L)\cdot\exp\left(-j(X)/2Z_L\right) \qquad (1)$$

The use of the transmitted wave instead of the reflected one reduces the error in $\underline{Z}$, specially in case of distributed impedance [1]. Taking the difference between object and reference measurement, as proposed earlier (1-5) induces additional errors due to asymmetries of the power splitter/combiner and non equal object and reference matching sections. A sequential measurement of the two lines allows the use of the same center conductor and matching sections. If instead of the difference the complex ratio of these two transmission measurements is computed, then this method is equivalent to a network analyzer procedure.

In a first set-up, the measurements are performed via a vector voltmeter (0.3 - 2000 MHz), an s-parameter test set (5 - 2000 MHz) and a synthesized signal generator controlled by a desk top computer following a flow diagram summarized in Fig. 2. Alternatively the HP 8510 system was often used having more comfort in operation but with a minimum frequency of 45 MHz being not very well adapted for certain objects and also showing some difficulties for special data treatment (k-parameter). In both cases, the frequency domain data were either evaluated directly in the frequency domain or converted into the time domain via FFT (or CHIRP-Z)) to obtain a synthetic pulse. Independent whether the raw data (= measurements) are taken in the F- or T-domain (sampling scope, not used here), a great advantage of time filtering consists in eliminating (by time window e.g. only the first pulse taken) the effect of multiple reflections on the final result. But it should be pointed out that the use of the time window has to be checked carefully since big errors or even meaningless results may be produced this way. The influence of the characteristic impedance, Z_L, has been systematically studied by using different wires (Z_L = 48Ω ≈ 50Ω matched, 125Ω resistively matched, 320Ω unmatched). Unavoidable small lengths differences between object and reference pipes were carefully measured mechanically and taken into account as a delay which may be approximated as an additional inductance. To achieve the high amplitude and phase stability required during a time to carry out reference and object measurement (≈ 30 min.), semirigid cables and APC 7 connectors were employed, and it was tried to keep the temperature variations in the laboratory as small as possible.

Longitudinal Measurements

A typical bellow (ID 100 mm. OD 120 mm, length 38 mm with 6 undulations) was measured in transmission and reflection (Fig. 3, Fig. 4) using time filtering. The beam impedance (imag. part) resulting from the transmission measurement was checked to be independent (within ± 10%) of Z_L. Good agreement was also found with theoretical results obtained by analytical formulation and by a numerical approach [2]. The apparent energy loss indicated in Fig. 3 by the decrease of $|S_{21}|$ (0.02 db at 1 GHz) is in fact explained by the measured reflected power (solid line) Fig. 4.

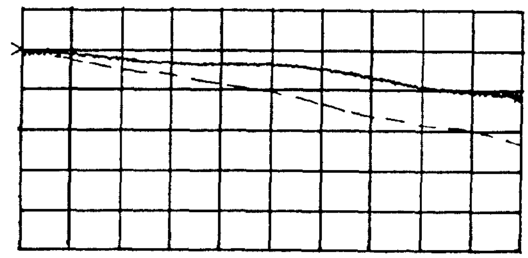

Fig. 3 Bellow Transmission Measurement
— Ampl. 0.02db/div; -- Phase 1°/div
45 MHz - 1000 MHz

If, as a check, $|S_{11}|$ resulting from the inductance L = 0.67 nH (derived from $\underline{S}_{21}$) is calculated one obtains much less reflection than measured (Fig.4). To construct a reasonable model which fits $\underline{S}_{11}$ and $\underline{S}_{21}$ and contains L = 0.67 nH, the effective ID of this bellow in the coaxial line was assumed to be reduced due to electric field concentrations on the corrugation. With a 48Ω "Wire" (48Ω because of mechanical ease) the bellow model consists of a 38 mm section of 42Ω line and L_S = 0.67 nH (Fig.4). This example shows that even for small, localised objects reflection measurements alone may give wrong results, and that it might be necessary to correct transmission measurements for the reflected energy. For the bellows mentioned-above the resonance frequency was found to be around 4.7 GHz. A distinct increase in ripple ($\underline{S}_{21}$, no time filter) above the beam-pipe cut-off frequency is attributed to waveguide mode interaction. Fig. 5 gives an idea for the reflected energy from a double sharp step (1 meter separation) and a double tapered step. The equivalent transmission (125Ω system) would consist of a 125Ω - 85Ω - 125Ω structure and a parallel capacitor at each impedance step (85Ω is a theoretical value for the elliptic beam-pipe with center conductor). With these impedance steps only, and time filtering provided to suppress multiple reflection one obtains (Z_{L1} = 125Ω, Z_{L2} = 85Ω) the transmitted power (Fig. 5, dashed line).

$$P_t/P_{inc}\approx 1-2\left((Z_{L1}-Z_{L2})/(Z_{L1}+Z_{L2})\right)^2= -0.36\ \text{db} \qquad (2)$$

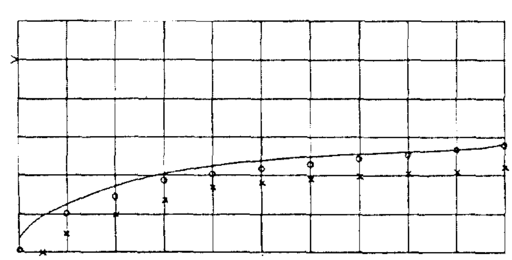

Fig. 4 Bellow Reflection Measurement
xxxx S_{11} from L=0.67 nH
oooo S_{11} from L=0.67 nH, Z_L=42Ω
10 dB/div; 45 MHz - 1000 MHz

To determine the beam impedance here only the phase-shift of the transmitted signal (not depicted) is relevant because Re{Z} is small. The tapered transition exhibits an interesting behaviour. For low frequencies the response is similar to the sharp step but for higher frequencies the taper acts like a matching section, thus reducing reflections. It should be noted that different time windows showed little effect on the response which might be considered as an indication for correct filtering. Also in this case a missing correction for the reflected energy would show up as high real part of $\underline{Z}$.

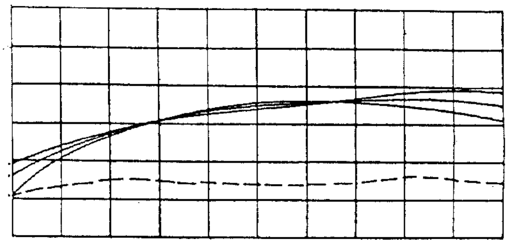

Fig. 5 Influence of Time Window (Gating); 2,3,4 ns
— Tapered Transition round-elliptic (2x)
-- Sharp Transition round-elliptic (2x)
45 MHz - 1000 MHz; 0.1 dB/div

Transverse Measurements

The transverse impedance could be obtained from measurements to extend the single coaxial wire technique towards a set-up with two parallel off- axis wires or a metallic plane in the center and one off-axis wire. Here the two-wire version was chosen, mainly because of mechanical simplicity. Since it would be rather difficult to provide broadband-matched end pieces for such a set-up it was decided to take simply flat end plates with N-connectors and to remove the influence of reflections by means of time-filtering. To ensure only odd mode excitation 180°, 3 dB power splitters (180° hybrids) were used. In this case an even mode rejection of at least 40 dB (20 dB either side) may be expected. Similar experiments with cavities (two off-axis probes) for mode selective excitation have shown a rejection of 60 to 80 dB for the unwanted mode. In the frequency range of interest (0.045-1GHz) the power splitter/combiner had no relevant amplitude ripple and phase distortions so that a relative clean pulse could be obtained.

Loss-Parameter

For k-Parameter measurements mostly the relation [3]

$$k = \frac{2Z_0 \int i_0(t)\left[i_0(t)-i_m(t)\right]dt}{\left[\int i_0(t)dt\right]^2} \qquad (3)$$

is applied under the condition that the pulse through the measurement object is time-shifted such, that the first 10 - 20% coincide with the reference pulse. Equation (3) is well adapted if real-time or sampling-scope techniques are used but problems may arise with synthetic pulse techniques. To generate a synthetic pulse where equation (3) can be applied, low pass interpolation of the FFT result is an implicit condition. This may create serious restrictions to the sampling process in the frequency domain, since in this case equidistant sampling, starting at $F = 0$ is required. Also a DC-component must be defined for cases where otherwise non tolerable truncation errors in the time-domain would occur. For bandpass-interpolation (HP 8510 system) the restriction of equidistant F-domain sampling is removed but the corresponding time-domain response is no longer directly comparable with a low-pass pulse. It has rather the meaning of the modulus of the complex envelope of a carrier modulated signal . Here it could be advantageous to use immediately the F-domain definition

$$k = \sum_n (\omega/4)\cdot(R/Q)_n \cdot \exp(-\omega_n^2 \sigma^2) \qquad (4)$$

For $Q \to \infty$ eq. (4) can be converted [5] into

$$k = \int \mathrm{Re}\,(Z) \cdot \exp(-\omega^2\sigma^2)\, d\omega \qquad (5)$$

The frequency range for practical k measurements is given by the frequencies where the constribution of the integrand becomes negligible. This is the case at low frequency where Re(Z) is still ≈ 0 and at high frequencies when $\exp(-\omega^2\sigma^2)$ approaches zero. It should be pointed out that eq. (5) is also approximately valid for finite Q.

Conclusion

As already pointed out the performed measurements show that correct results are most likely obtained in transmission (and not in reflection). This requires in general high stability and very good reproducibility (±0.01 dB, ± 0.5 deg at 1 Ghz) of the test bench including the network analyzer, cables and connectors, since the observable changes in amplitude and phase may be small. Time filtering is an effective tool to remove multiple reflections. Measured results in terms of beam-impedance of various $|Z| \ll Z_L$ (bellows, pickup, steps) for different characteristic impedances (50 Ω, 125 Ω, 320 Ω) are practically independent of the center wire diameter and agree relatively well with theoretical values. Care must be taken not to excite higher order modes if the measurement frequencies are above cut-off for the beam-pipe, particularly if no time-filtering or k-parameters evaluations are used.

Acknowledgements

The author would like to thank J.P. Delahaye, G. Dôme, H. Henke and F. Pedersen for very valuable discussions as well as D. Boussard, J.H.B. Madsen and E. Jones for their support and finally B. Szeless for the mechanical work.

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