Minimizing crosstalk in a high-speed cable-connector assembly.

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Abstract

This paper presents the detailed signal-integrity analysis results of the connector and cable linking the ALICE Time Projection Chamber (TPC) to its Front-End Electronics.

The goal was to find a cable design that minimizes the crosstalk (electromagnetic coupling) between the signal lines. Other considerations taken into account were the signal line capacitances and the imposed mechanical constraints (cable flexibility, thickness, and physical dimensions).

Crosstalk effects in the connector pins were also analysed.

The design was tackled using different software tools. For the cable, a Finite Element Method was used to extract an equivalent distributed circuit model which was then exported to PSpice®. The resulting simulations will be presented. For the connector, an electromagnetic full-wave solver was used to simulate completely all high-speed effects.

We will show how these programs helped us to quickly investigate different cable configurations.

I. PROBLEM DESCRIPTION

The ALICE TPC detector has to be linked to the read-out electronics by some electrical means. The most convenient way of doing this is by a flexible flat cable. However, the nature of the signals to be transmitted (low-amplitude, analogue, very fast rise-times) requires a thorough study of the characteristics of this transmission line, particularly regarding signal integrity issues such as crosstalk and reflections.

The main objective of this design is to minimize crosstalk induced by a signal pulse on the neighbouring traces to avoid false triggering.

Several factors influence crosstalk, but the most important are signal rise-time, coupling length, track separation (i.e. signal pitch), stackup and the form of the propagating electromagnetic fields. Given the problem constraints, we will focus on how to bound the electromagnetic fields so as to reduce coupling between channels. The only method practically available for achieving this is by proper ground and signal placement in the cable and connector.

There are some other effects to consider. First, the capacitance to ground has to be kept low so as to reduce the parasitic effect to the charge detector input. Also, the cable has to remain flexible, so that a proper mechanical connection and maintenance is ensured. This implies that we need a minimum thickness cable with no solid planes.

The cable is built on a flexible PCB with one layer of polyamide (ε_r=3.5) of 70 μm thickness between two 55 μm thick layers of coverlay (ε_r=4.4). The copper signal tracks are 160 μm wide and are separated by 640 μm.

The cable will be mated to a PCB using a connector. This will play an important role in the system behaviour due to the fast (~50 ps) edge rates and so has to be properly characterised.

II. CROSSTALK DEFINITIONS

Crosstalk occurs in any system with two or more conductors. Each wire segment acts individually as an inductor and capacitor, and also as an antenna. Together, they act as coupled antennae due to their mutual coupling. This coupling can be expressed in terms of capacitive and inductive components. Figure 1 shows a simple crosstalk scenario with one active (aggressor) and one passive (victim) line.

The nodes of the transmission lines are labelled:

1. Near-end (driver-end) of the aggressor line
2. Far-end of the aggressor line
3. Near-end of the victim line (backward crosstalk)
4. Far-end of the victim line (forward crosstalk)

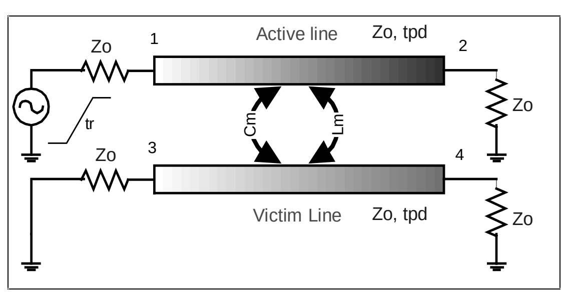

Figure 1: Crosstalk scenario. C_m models the capacitive crosstalk; L_m models the inductive one.

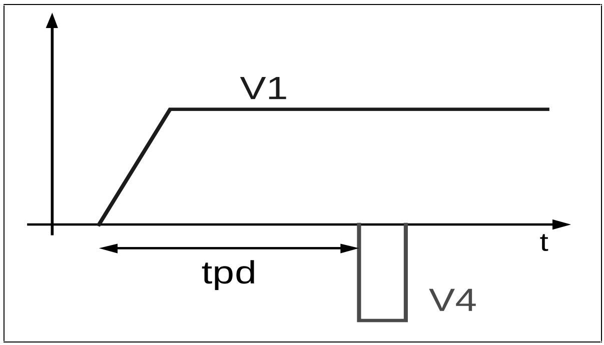

Figure 2: Forward crosstalk (V4) in the far-end of the victim line for an aggressor signal (V1).

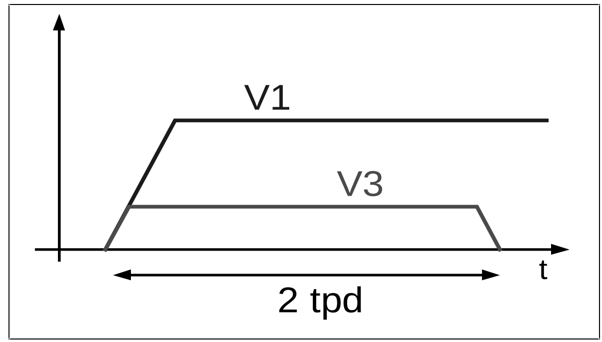

Figure 3: Backward crosstalk (V3) in the near-end of the victim line for an aggressor signal (V1).

Note that matched terminations are assumed on all ports. The coupling parameters (Cm, Lm) are given for the coupled length.

Figure 2 shows the theoretical signals for the far-end crosstalk for this configuration.

The amplitude of the induced signal can be evaluated as [1]:

$$K_f = \frac{1}{4}\left(\frac{C_m}{C} - \frac{L_m}{L}\right) \qquad \text{(Eq. 1)}$$

$$V_4 = V_1 K_f \left(\frac{2t_{pd}}{t_r}\right) \qquad \text{(Eq. 2)}$$

The forward crosstalk coefficient is proportional to the difference of the normalised capacitive and inductive couplings. If both components were equal, there would not be any far-end signal. This is known as homogeneous propagation. However, under typical non-homogeneous circumstances, the inductive component is larger than the capacitive component and the overall crosstalk has a negative polarity. Its magnitude is proportional to the coupled length and inversely proportional to the signal rise-time and the track separation.

Figure 3 shows the waveforms for the near-end crosstalk.

Here, the crosstalk coefficient is given by [1]:

$$K_b = \frac{1}{4}\left(\frac{C_m}{C} + \frac{L_m}{L}\right) \qquad \text{(Eq. 3)}$$

$$V_3 = V_1 K_b \qquad \text{(Eq. 4)}$$

The backward coefficient is proportional to the sum of the normalised capacitive and inductive couplings, and so it can never be zero or negative. The backward crosstalk will always have the same polarity as the aggressor signal and its duration will be equivalent to twice the propagation delay of the traces.

Due to the small cable-thickness, the electromagnetic fields will propagate in both polyamide and air. This leads to a non-homogeneous propagating structure and so to both forward and backward crosstalk.

III. MINIMIZING CROSSTALK IN A PCB

A. *Simulation Tools*

Assuming that all fields in the cable will be transversal to the direction of propagation, we can analyse the structure using the Maxwell 2D Extractor®. This is a quasi-static solver that leads to solutions where the high frequency effects (i.e. radiation) are ignored. This approach is valid as long as the physical dimensions are smaller than one tenth of the wavelength considered.

The methodology consists in initially drawing a 2D model for the cable. The signal lines and signal returns (ground) have to be identified, the cable materials declared and finally, the problem boundary conditions defined.

Maxwell 2D Extractor® is then used to provide the electromagnetic fields solution for the structure. The program outputs the capacitance and inductance matrices for the problem (computed by integrating the fields over the cable) and can also extract parameters such as characteristic impedance, crosstalk coefficients and the propagation delay.

The inductance and capacitance matrices represent an equivalent transmission line model that can be exported to a SPICE simulator. This allows us to simulate the structure as part of a larger electrical circuit and obtain the resultant signals in the time domain.

The Maxwell 2D Extractor® can create either lumped or distributed models. Lumped models can include losses and frequency dependency but can produce spurious results. Distributed models are generally better behaved but they cannot include losses or frequency dependent effects. It is up to the designer to choose the most suitable representation for a given problem.

Another important feature of this tool is the parametric sweep option that was used to find the optimum solution. The dimensions of the structure can be varied so as to find the configuration with the lowest possible capacitance and coupling. This was accomplished by running multiple simulations on the several possible physical structures.

B. Simulations

The 2D simulations are accurate and computationally inexpensive, so many different layouts could be quickly tested (Figure 4).

The necessary cable flexibility leads to models with the ground plane split into several tracks. The first two models were used to study the effect of guard tracks set between the signal tracks. In models C and D, the signal tracks have been placed on opposite sides of the polyamide layer in order to increase the effective pitch and thus hopefully reduce crosstalk.

Distributed lossless SPICE equivalent models were extracted. This was considered a good approximation as losses were estimated as negligible over the small cable length (75mm).

The models were exported to PSpice® to check the crosstalk result in the time domain. A gaussian pulse of 50ps rise time (fastest expected pulse case) was applied to the aggressor track. Except for the near-end of the aggressor, all the cable ends were terminated in the characteristic impedance of the structure. The crosstalk was measured in all of them.

C. Results

For the cable design, the low coupling and the low self-capacitance requirements are contradictory. If the signal return paths are very close to the signals themselves, the electromagnetic fields will be mostly contained in this area and thus there will be minimal stray fields and coupling with other lines. However, this implies a larger capacitance to ground with its adverse consequences to the charge detector.

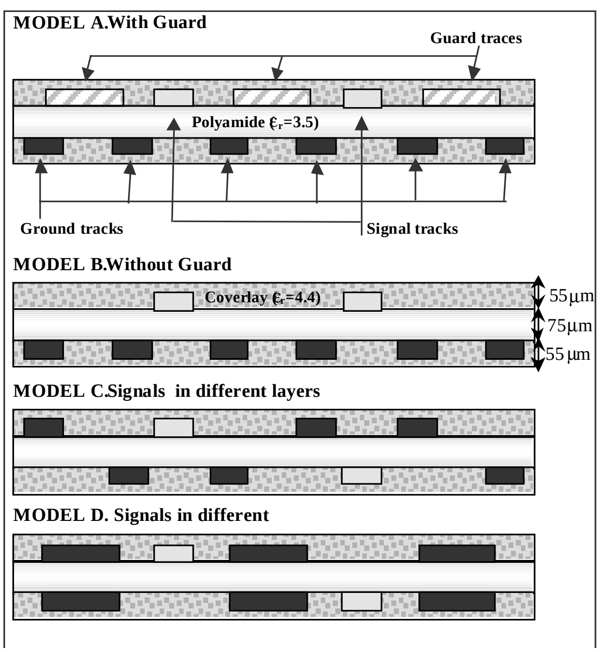

Figure 4: Different signal-ground track configurations.

The total cable and connector-assembly capacitance should not exceed 10pF so as to limit the parasitic effects to the input charge detector. The calculated cable capacitance is well below this limit (table 1).

Table 2 shows the good correlation between the simulated crosstalk values in PSpice® and the calculated theoretical values (equations 1-4). We see how the guard tracks used in model 'A*' can help to reduce considerably the crosstalk. However, as both ends are terminated to ground, a standing waveform can arise and produce undesirable resonances [2]. This can be avoided by grounding the lines at one of the ends through termination resistors (model 'A**') but this also much reduces the guard track effect (Table 2).

Figure 5 shows the PSpice® simulation results for every model. The black traces correspond to the near-end crosstalk, the grey ones to the far-end crosstalk. The results for model 'A*' clearly show the effects due to the line resonances - these are much reduced for the 'A**' configuration.

The bottom traces show the estimated crosstalk for models B and C (model D showed similar behaviour but with peak values lying between those of B and C).

Model C was finally selected due to its considerably lower crosstalk.

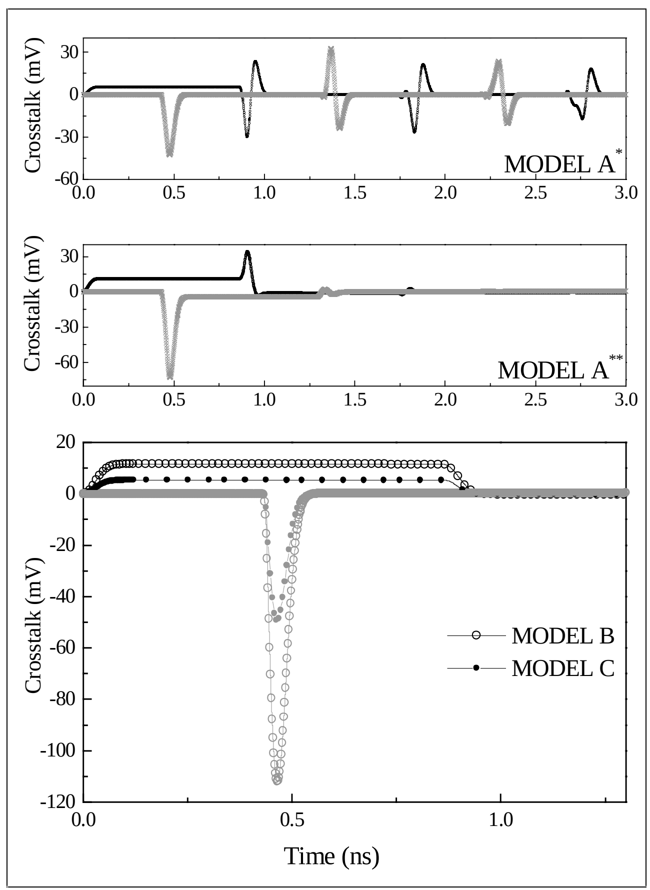

Figure 5: Far-end and near-end crosstalk for the different configurations.

Table 1: Characteristic parameters for different configurations

Model Name	C (pF)	C_m (pF)	L (nH)	L_m (nH)	Z_0 (Ω)	t_{PD} (ps)
A	7.73	0.04	25.3	1.0	57.26	442.8
B	7.25	0.08	26.0	0.9	59.6	436.5
C	7.36	0.04	25.8	0.4	59.2	436
D	5.99	0.05	28.8	0.7	69.3	415

Table 2: Forward and backward crosstalk for a 50 ps rise-time aggressor signal vs. theoretical values.

Model Name	V4 (mV) Forward	V3 (mV) Backward	V4 Theor. (mV) Forward	V3 Theor. (mV) Backward
A*	-42.5	5.3	-	-
A**	-72.4	11.1	-	-
B	-111.8	11.5	-102.9	11.4
C	-48.9	5.3	-43.9	5.2
D	-62.5	8.0	-66.2	8.2

* Guard tracks to ground.

** Guard tracks loaded with Z_0 at the far-end.

IV. ANALYSIS OF THE CONNECTOR CROSSTALK

D. Tools

Due to the high-speed nature of the application, SAMTEC's 0,80 mm hi-speed (QSE and QTE 020-01-L-D) connectors were chosen. As these are commercially bought items, the only practical option available to the designer to reduce crosstalk is by strategic signal and ground pin assignments.

The complex structure of the connector and the very fast edge rates expected requires an analysis with a full wave solver. This allows us to take into account effects such as radiation.

The simulation procedure includes the creation of a 3D CAD physical connector model, the specification of the materials, the identification of signals and signal return paths, the port definitions and the input excitation description.

Since MicroWave Studio® uses the Finite Integration Technique (FIT), it is able to show the resultant signal waves directly in the time domain. This allows the study of possible signal integrity issues (crosstalk, ringing, reflections etc) without the need of a SPICE simulation. It also provides the S-parameters of every port for a defined frequency range and the electromagnetic field distribution at any given frequency.

The gaussian input pulse used is the ideal excitation signal for time domain simulations as its transformed frequency domain signal is also gaussian with a limited bandwidth and no zero axis crossings. This simplifies the S-Parameter calculations.

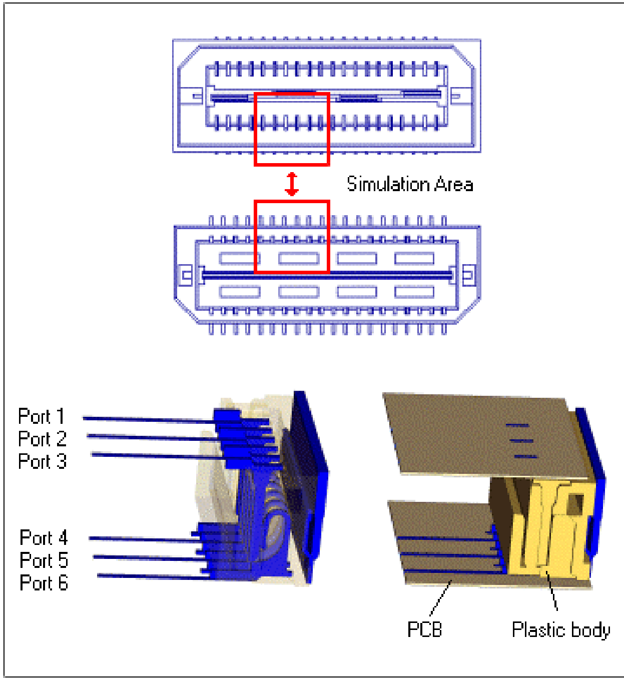

Figure 6: 3D connector model.

E. Simulations

We will present the crosstalk characteristics for a given signal and ground pin configuration

We exploited the high level of symmetry in the connector so that we needed only simulate one part of it (Figure 6). This alleviated the meshing problems and allowed model simulations in reasonable times.

The second snapshot in Figure 6 represents the metallic parts of the mated male and female connectors. The third figure shows the same mated connectors but now with the outer plastic bodies and already connected to two small pieces of PCB.

An alternating ground-signal-ground configuration was used in order to minimize the crosstalk.

A 50 ps rise-time gaussian pulse was applied to port 2 (aggressor). The near-end crosstalk was measured using ports 1 and 3, the far-end crosstalk at ports 4 and 6.

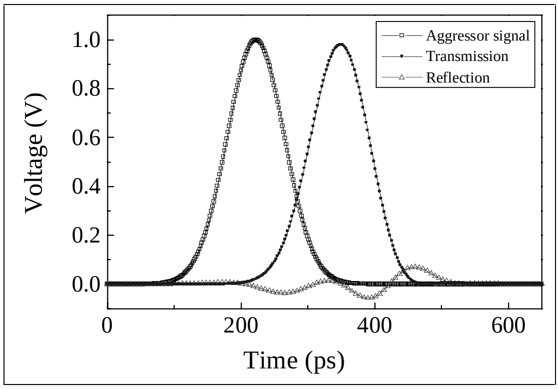

Figure 7: Time-domain connector transmission results.

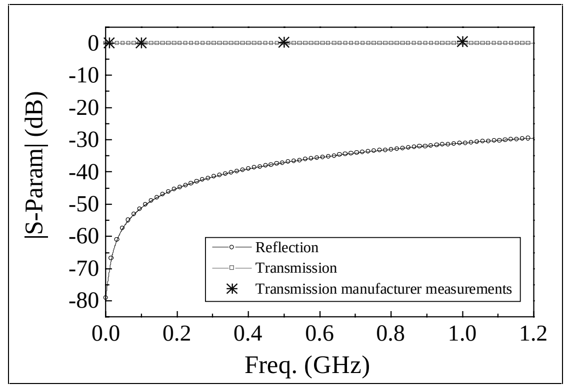

Figure 9: Frequency-domain connector transmission results.

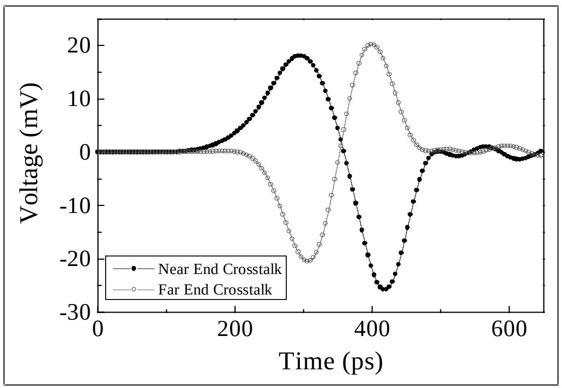

Figure 8: Near-end (port 1) and far-end (port 4) crosstalk

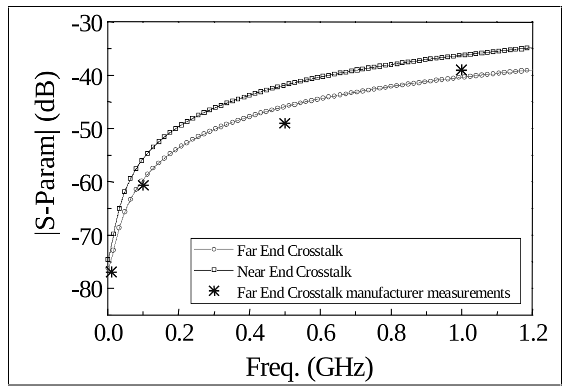

Figure 10: Near-end ($|S_{21}|$) and far-end ($|S_{24}|$) crosstalk

F. Results

Figure 7 shows the transmitted and reflected signals for the given input. Almost all of the input energy is transmitted to the output with very little reflected back towards the source. Figure 8 shows the near-end crosstalk for port 1 and the far-end crosstalk for port 4.

Figures 9 and 10 show the simulated S-parameters. The star symbols show measurements provided by the manufacturer for a similar signal-ground pin configuration [3]. There is good agreement between the calculated and measured results.

V. CONCLUSIONS

It has been shown how signal-integrity issues become important with the use of high-speed signals. We have demonstrated how crosstalk can be simulated and what measures can be taken to minimize it.

For the cable design, it was practical to simulate different configurations until an optimum solution was found. The final equivalent distributed model obtained from Maxwell 2D Extractor® was exported to PSpice® and simulated as part of a larger electrical circuit.

For the connector analysis, the MicroWave Studio® full-wave solver was used due to the fast signal rise-time and small connector dimensions involved.

The good correlation between our simulations and those predicted by theory and from the manufacturers data allows us to have confidence in our results.

We have demonstrated how these tools can be used to replace ad-hoc and rule of thumb design approximations. They can help to minimize the number of possible design iterations and can greatly aid in completing projects within time, budget and personnel constraints.

All of the programs discussed in this paper are available and fully supported at CERN by IT/CE-AE.

VI. ACKNOWLEDGMENTS

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VII. REFERENCES

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[4] Maxwell 2D Extractor® manual.

[5] CST MicroWave Studio® manual.