COUPLING OF FAST TRANSIENTS FROM POWER SUPPLY LINES INTO COAXIAL LINES: EXPERIMENTAL AND THEORETICAL RESULTS

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1. Introduction

The problem to be discussed in this paper has been studied in connection with signal transmission in a digital loop communication system [1]. We had to investigate how to predict disturbances caused by fast transients that are present on nearby power supply lines. The aim of our work was not so much to get precise quantitative results, but to understand qualitatively the coupling mechanisms and effects, and to estimate the maximum possible coupling for practical worst case situations. We tried to simplify the theoretical models as much as possible, as long as they still allowed us to understand the principal effects.

2. Noise Sources of Fast Transients

Every electronic engineer knows that fast transients may exist in the power distribution system. Potential sources of these transients are mainly mechanical switching devices, in other words, everywhere electric arcing occurs. Up to now, there has been no standard test pulse defined, but we hope in future it will be possible to agree on a standardization. From [2, 3] as well as from our own measurements we find that a risetime between 5 and 20 ns and an ideal voltage source between 1000 and 1500V has to be assumed. The voltage measured in the power supply distribution system will for most cases be below 500V.

3. Coupling from Power Supply Lines to the Coaxial Shield

3.1 Multiconductor Models

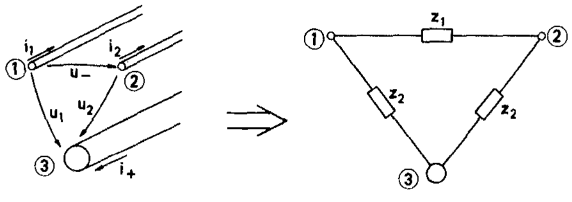
To predict the transient electromagnetic wave on the outside surface of the coaxial line we can use some results of the theory of lossless multiconductor systems. A clear description of the fundamental relations in lossless multiconductor systems with TEM (Transversal electromagnetic wave) are given in [4, 5]. The general model assumes a lossless parallel arrangement of n conductors with a constant cross sectional geometry along the section length in consideration. The dimensions of the cross sectional geometry have to be an order of magnitude smaller than the section length l, leading to a dominance the TEM. The dielectric may be inhomogenuos over the cross sectional geometry.

The multiconductor theory gives the folowing results:

- The n-conductor system can be represented by n characteristic impendances and (n-1) transmission modes.
- One mode consists in general of a linear combination of voltages between all conductors.
- Each mode may have its own propagation velocity. In the case of a homogenous dielectric, all modes have the same propagation velocity.
- Coupling between the modes as well as reflection can only occur at the beginning and at the end of the loosless multiconductor system. Along the legnth of the multiconductor system, the modes are propagated unchanged.
- The mode coupling and reflection at the ends of the multiconductor system depends on the relationship between the characteristic impedances of the multiconductor system and the attached impedances at the ends (similar to the reflection coefficients in 2 conductor systems).

The simplest multiconductor model is the symmetric three conductor model.

The definitions are taken from [5] where the model has been used for prediction of the coupling in printed circuits.



$u_+ = u_1 + u_2$ $\qquad$ $u_- = u_1 - u_2$

$i_+ = i_1 + i_2$ $\qquad$ $i_- = i_1 - i_2$

$z_+ = \frac{u_+}{i_+} = \frac{u_1 + u_2}{i_1 + i_2}$ $\qquad$ $z_- = \frac{u_-}{i_-} = \frac{u_1 - u_2}{i_1 - i_2}$

z_1, z_2: impedances between the conductors

$z_1 = 2\,\frac{z_+ z_-}{z_+ - z_-}$ $\qquad$ $z_2 = z_+$

u_-, i_-: „series mode wave"

u_+ i_+: „common mode wave"

Fig. 1: Definitions of the symmetric 3-conductor model

u_1, u_2, and i_1, i_2 are the voltages and currents as they are measured at conductors 1 and 2, taking conductor 3 as a reference. u_+, i_+ and u_-, i_- are the two modes of this three conductor systems. In this special case, the modes can be identified in the circuit: u_- can be measured as the voltage difference between conductors 1 and 2 and i_+ can be measured as the current in the reference conductor. (This direct identification of the modes is only possible in the symmetric three conductor system).

The impedances of real multiconductor systems can be measured with the "Time Domain Reflectometer" (TDR). The TDR always measures a combination of impedances, but individual values can be calculated from these measurements.

Fig. 2 shows the measured voltages at the end of a 3-conductor system consisting of 2 wires in a power supply line and the shield of a coaxial line as the reference conductor, being in parallel over a length of 95 m. u_1 and u_2 contain a mixture of the two modes. Summing and substracting the two signals gives the original mode voltages. u_- is propagating slower than u_+ because of the differences in the dielectrica.

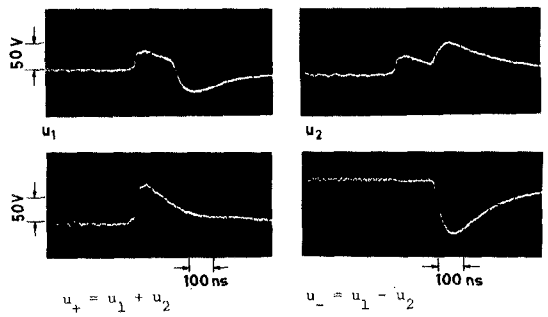


Fig. 2: Measured signals at the end of a multiconductor system with power supply line and coaxial cable.

3.2 Coupling Planes

To adapt this model to practical situations we have to find out all the places along the propagation path of the transients where the cross sectional geometry is changing. At all these places, which we call "coupling planes", it is possible to estimate the coupling according to the impedance relations in the coupling plane. As an example of a coupling plane Fig. 3 shows the junction of a 2-conductor system to a 3-conductor system, and the equivalent circuit in the coupling plane. R_1 and R_2 are additional impedances that may be present between the conductors in the coupling plane.

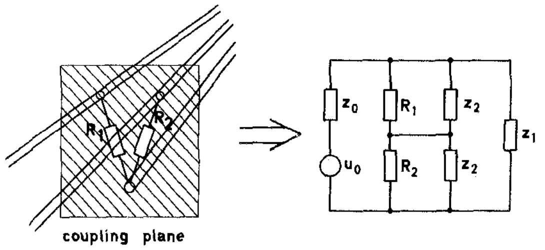


Fig. 3: Example of a coupling plane

From the equivalent circuit we can calculate the general coupling relations for this type of coupling plane:

$$u_- = u_0 \frac{2Z_1 + Z_1 Z_2 \frac{R_1 + R_2}{R_1 \cdot R_2}}{Z_0\left(2 + \frac{Z_1}{Z_2}\right) + 2Z_1 + \frac{Z_0 Z_1 Z_2}{R_1 R_2} + (Z_0 Z_1 + Z_0 Z_2 + Z_1 Z_2)\,\frac{R_1 + R_2}{R_1 R_2}}$$

$$u_+ = u_- \frac{Z_2 (R_1 - R_2)}{2 \cdot R_1 R_2 + Z_2 (R_1 + R_2)}$$

This general coupling relation is simplified if one of the following conditions exist:

1. If R_1 and R_2 are infinite we get:

$$u_- = u_0 \frac{1}{1 + \frac{Z_0}{Z_1} + \frac{Z_0}{2Z_2}} \qquad u_+ = 0$$

2. If R_1 and R_2 are equal we get:

$$u_- = u_0 \frac{1}{1 + \frac{Z_0}{Z_1} + \frac{Z_0}{2Z_2} + \frac{Z_0}{2R}} \qquad u_+ = 0$$

3. If R_1 is infinite and R_2 is zero, we get:

$$u_- = u_0 \frac{1}{1 + \frac{Z_0}{Z_1} + \frac{Z_0}{Z_2}} \qquad u_+ = u_-$$

Similar calculations can be made for other types of coupling planes, but many real situations can be interpreted by this simple coupling plane model.

3.3 The Wave at the Outside Surface of the Coaxial Shield

When the coaxial shield represents the reference conductor of a symmetric 3-conductor system, the wave at the outside surface of the coaxial shield is:

$$i_c = i_+ \quad u_c = \frac{u_+}{2} \quad Z_c = \frac{Z_+}{2} = \frac{Z_2}{2}$$

Calculations and measurements of different possible situations show that with the sources specified in Sec. 2, i_c is below a few Amp. and u_c is in general less than 500V. How far the wave propagates in the power supply distribution system, depends on the details of the wiring. Every irregularity in the wiring yields to reflection and coupling of the original modes, which lead to a rapid distortion of the transients. There may, however, exist special situations, where one of the modes is propagated as a TEM wave over a longer distance. In this case damping, because of dielectric losses and small geometry fluctuations, is dominating. Our experience shows a damping of about a factor of 2 over a length of 100 m with the test signals we were using. (risetime $\approx$ 10 ns, decay time $\approx$ 100 ns).The twisting of the power supply cable has not much influence on the propagation of this mode.

4. Coupling into the Coaxial Line

To determine the coupling of the wave outside the coaxial cable into the signale circuit, we have to distinguish between:
- coupling along the cable
- coupling at the cable end.

4.1 Coupling Along the Cable

The coupling along the cable can be calculated by using the concept of transfer impedance and transfer admittance [6, 7].

4.1.1 Directional Coupling Effect

In the case of a parallel arrangement of a power supply line and a coaxial line between points A and B with the wave u_c, i_c and the impedance Z_c, we will observe a "directional coupling" into the coaxial line. Each differential segment acts as a local source of electromagnetic energy (depending on the transfer parameters, and proportional to the length dx of the differential segment). The signal of this source is propagated in both directions inside the coaxial cable. In one direction the internal signal is propagated together with the exterior noise source, in the other direction it is not. Only in one direction the contributions of all differential segments will be summed up and result in a measurable voltage transient inside the coaxial line.

4.1.2 Transfer Parameters

The coupling into a differential segment dx of the coaxial cable can be calculated with transfer impedance and transfer admittance. Taking the definitions of Vance [7] we will have a distributed voltage source of

$$du = i_{c_{(t)}} \cdot Z_T \cdot dx$$

and a distributed current source

$$di = u_{c_{(t)}} \cdot Y_T \cdot dx$$

Z_T is the transfer impedance (per unit length) Y_T the transfer admittance (per unit length)

$$Z_T = Z_D + j\omega M_{12}$$

$$Y_T = -j\omega C_{12}$$

Z_D is the diffusion term of the transfer impedance

M_{12} is the inductive coupling term of the transfer impedance

C_{12} is the capacitive coupling term of the transfer admittance.

For a cable with impedance Z_0 the contribution of these two terms on the internal wave propagating in the same direction as the exterior wave can be combined:

$$\frac{du}{dx} = \frac{1}{2} \cdot i_c \left[\underbrace{Z_d}_{\text{diffusion term}} + \underbrace{j\omega (M_{12} - Z_0 Z_c C_{12})}_{\text{penetration term}} \right]$$

Z_0: Impedance of the coaxial cable.
Z_c: Impedance of the exterior system.

For the transfer coupling in a given configuration, we call the first term the diffusion term, and the second term the penetration term.

The diffusion term is representing the diffusion of the current i_c through the shield. Z_D is frequency dependent, and approaches the series resistance R_0 of the casble shield for low frequencies and is smaller than R_0 [6, 7] at higher frequencies.

The penetration term represents the penetration of the external electromagnetic field through the small apertures of the braided shield. It may be positive or negative, depending on the relative contribution of capacitive and inductive coupling. For a given configuration, we may represent the penetration term with one single mutual inductance:

$M_p = M_{12} - Z_0 Z_c \cdot C_{12}$

To simplify the worst case calculations we may take Z_d = const = R_0 and we are then able to calculate the response in the time domain with

$$\frac{du_{(t)}}{dx} = \frac{1}{2} i_{(t)} Z_d + \frac{d}{dt} i_{(t)} M_p$$

where $du_{(t)}$ is the voltage wave, propagating inside the coaxial line in the same direction as the source wave outside from a differential increment dx of the coaxial line.

The voltage at the end of the coaxial line is the sum of all these incremental contributions.

4.1.3 Dispersion Because of Velocity Differences

Because of the difference in the propagation velocity of internal and external waves in the coaxial cable, a transient with a very short risetime will not cause a signal of the same risetime inside the coaxial line. The effect of this dispersion can be treated separately for the diffusion term and the penetration term.

4.1.3.1 Dispersion of the Diffusion Term

As an example we calculated the dispersion of an exponential pulse with very short risetime and a decay time of

$i_c = 0$ for $t < 0$

$i_c = i_0 \cdot e^{-\alpha t}$ for $t > 0$

V_1 is the propagation velocity of the exterior wave i_c

V_2 is the propagation velocity inside the coaxial cable

L is the length of the coupling section

integrating the contribution of all differential segments to the voltage at the end of the coupling section gives the signal u_L as shown in Fig. 4.

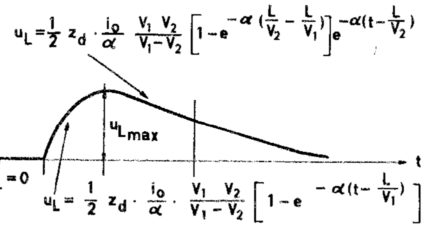


Fig. 4: Signal u_L resulting at the cable end from diffusion term

The maximum is at $t = \frac{L}{V_2}$ and is

$$u_{max} = \frac{1}{2} Z_d \frac{i_0}{\alpha} \frac{V_1 V_2}{V_1 - V_2} \left[1 - e^{-\alpha L (\frac{1}{V_2} - \frac{1}{V_1})}\right]$$

The time between the beginning of the pulse and the maximum is

$$\Delta t = L\left(\frac{1}{V_2} - \frac{1}{V_1}\right) \qquad \text{for } \frac{1}{\alpha} \gg \frac{L}{\frac{1}{V_2} - \frac{1}{V_1}}$$

The maximum will be less than

$$u_{max} < \frac{1}{2} R_0 \cdot i_0 \cdot L$$

The calculation with $Z_T = R_0$ = const is valid if

$$\frac{1}{\alpha} \gg \Delta t > \tau_s$$

τ_s is the diffusion time constant of the shield [7].

If this condition is not true, then the coupled signal will be less than this calculation. This calculation is a worst case estimation.

4.1.3.2 Dispersion of the Penetration Term

As we have seen in 4.1.2, we may combine the inductive and capacitive penetration coupling to a combined penetration coupling term M_p. The signal resulting from this term, for a differential segment dx in the time domain, is the derivative of the external signal:

$$du = \frac{d}{dt} i_{c(t)} \cdot M_p \cdot dx$$

The calculation is simplified if we replace the sharp edge of the fast signal in 4.1.3.1 with a short but finite rise time, and let the decay time be very long.

$i_{(t)} =$ for $t < 0$

$i_{(t)} = \frac{i_0}{t_0} \cdot t$ for $0 < t < t_0$

$i_{(t)} = i_0$ for $t_0 < t$

Integrating the contribution of all differential segments to the voltage at the end gives a signal shown in Fig. 5.

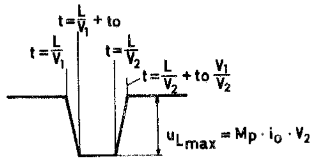


Fig. 5: Signal u_L resulting from penetration term

The condition for this pulseform is:

$$t_0 < \frac{L}{V_2} - \frac{L}{V_1}$$

The maximum voltage at the cable end is

$$u_{max} = M_p \cdot i_0 \cdot V_2$$

If the condition is not satisfied the maximum of the coupled signal will be less. This maximum is a worst case estimation.

It is interesting to note that coupling from penetration term, is independent of the section length, for longer distances.

Fig. 6 shows two measurements of a signal u_L at the end of a coupling section of 50 m and 100 m length, with all other conditions (geometry, test signal, impedances, etc.) the same. The positive peaks result from penetration coupling and are practically the same for both the 50 m and 100 m section lengths. The negative and much slower peaks result from diffusion coupling, and are proportional to the cable length.

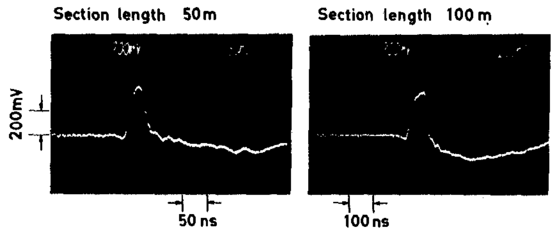


Fig. 6: Measurement of u_L with two section lengths.

4.2 Coupling at the Cable End

The coupling at the cable end depends to a high degree on the details of the construction at the coaxial connection. It is beyond the scope of this paper to discuss this problem in detail. Some short remarks have to be enough:

- If the shield is attached with even a short piece of wire, this inductivity will result in a very high transfer impedance at this place.

- If the coaxial shield is not connected, the common mode voltage u_c will be divided according to the impedances and the coupling be very high.

- If a galvanic separation of the line circuits is intended, the input region of the circuit must be very carefully designed, and has to be tested with voltage transients in the heigth of the expected u_c voltage between shield and local ground.

5. Measurement Method for Cable Qualification

In the literature, [6, 7] measurement of cable transfer characteristics in a triaxial arrangement is described, where the cable sample has usually to be electrically short. We did not have an appropriate triaxial test chamber and we know, that proper connection of the cable ends in such a chamber to avoid faulty measurements is a problem. We also think that measurement in a relatively low impedance test chamber could underestimate the contribution of the capacitive coupling term.

We therefore tried another approach:

- We used a time domain test signal (similar to the signals expected in a noisy environment).

- We used an electrically long test sample (compared to the rise and fall times of the test signal).

- We used a test configuration similar to real coupling situations.

As a signal source we used an interference generator as it is used for simulation of fast transients in the power supply.

The signal coupled into the cable sample was measured with a fast storage oscilloscope.

The test configuration was a plastic conduit with a fixed power supply cable as a noise source and a channel for the cable under test in the distance of 25 mm to the power supply cable. The plastic conduit was mounted in form of an U, so that the cable for triggering the oscilloscope was shorter than the test sample. The layout of the supply cable and the trigger cable to the oscilloscope had to be optimized carefully to avoid direct coupling into the oscilloscope. The coaxial cable under test was grounded at both ends and was acting as a HF-Ground for the multiconductor system.

The impedances and the velocities were measured with a reflectometer, and the multiconductor system was matched with appropriate resistors.

The sample length was about 30 m. This was a good compromise for our equipment: The sensitivity on rise time is reduced, the damping of the u_c wave is less than 0.8, different parts of reflections can be easily identified and handling of the cable samples is still practical. The conditions for measuring the maximum of the penetration contribution (4.1.3.2) is met.

Fig. 7 shows some results obtained in this test configuration.

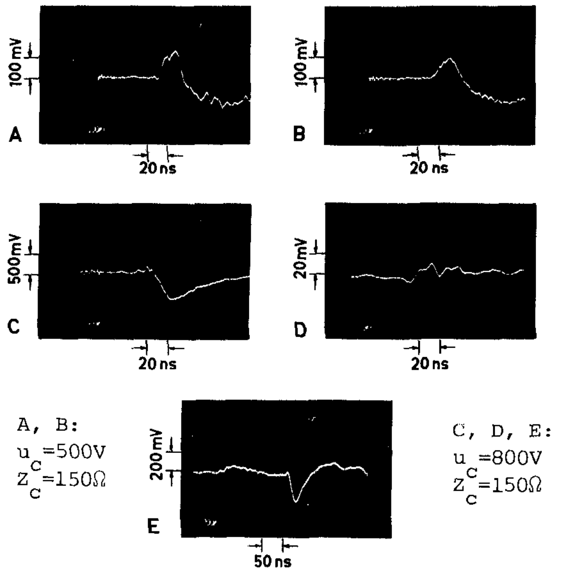


Fig. 7: Measurements in the 30 m cable test configuration.

Measurement A and B show the signal in a single braid cable similar to the popular RG59/U type. The risetime of the test signal was changed from 5 ns in measurement A, to 30 ns in measurement B. This example demonstrates the insensitivity of the measurement method to the risetime of the test signal.

Measurements C is the signal in a cable with AL-foil and a very light Cu-braid, intended for use in CATV distribution. In this case, diffusion coupling is dominating.

Measurement D is the signal in a double braid cable (without isolation between the braids). The coupling is much less and it is not possible to separate the relative contribution of the two coupling terms. Similar results are obtained with a triaxial cable, when both ends of both shields are perfectly grounded.

Measurement E shows the coupling into a triaxial cable, when the outer shield is grounded at one end only. The result is even slightly worse than with a single braid shield cable and can be explained easily with the theoretical models of this paper: At the open end of the outer shield, the u_c wave is coupled into the mode between the shields according to the impedance relations. From here, a reflected wave is travelling back between the shields, resulting to a directional coupling into the signal path. As the impedance here between the shields is very low and the travelling speed for the wave between the shields is also lower than for the wave between a single braid shield and power supply line, the diffusion term is dominating in this case.

6. Conclusions:

The models used in this paper are useful to understand qualitatively the propagation and coupling mechanisms of fast transients and to estimate the worst case values. For practical worst case calculations we propose to use a DC measurement of R_0 for the diffusion term, and the measurement according to section 5 for the penetration term.

The measurement method of section 5 is published as a basis for further work. We think, it will still be possible to optimize this measuring configuration, to get more precise results.

Using a configuration with air-dielectric and a very straight single copper wire would give more speed difference and less dispersion of the external wave. Together with a signal source of very short rise time and a properly matched configuration, this would allow the use of shorter sample lengths.

7. References

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