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Y-J. Chen
G. Caporaso

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DESIGNS FOR A HIGH POWER SUPERCONDUCTING DELAY LINE

Yu Ju (Judy) Chen and G. Caporaso, Lawrence Livermore National Laboratory, University of California, P.O. Box 808, Livermore, CA 94550

Abstract

Potential designs for a high power superconducting delay line of approximately 10 μ s duration are described. The transmitted signal should have low dispersion and little attenuation to recapture the original signal. Such demands cannot be met using conventional metal conductors. This paper outlines a proposal for a new transmission line design using low temperature superconducting material which meets system specifications. The 25 Ω line is designed to carry pulsed signals with an approximate rise time of 8nsec and a maximum voltage magnitude of 25kV. Predicted electrical design and performance of the line will be presented.

1 APPLICATION OF THE LINE

The line will be used to provide transit time isolation of the reflections expected to be generated in the kicker of a multiple line-of-sight pulsed accelerator[1]. The kicker will operate with beam currents of 3 to 6 kA and will launch beam-induced voltages up the transmissions lines connecting it to its pulsers. The most straightforward way to eliminate the effects of these reflections is to transit time isolate the kicker from the pulser. Thus the 10 μ sec line will actually allow a 20 μ sec operational interval.

2 WHY USE SUPERCONDUCTORS?

A 10 μ s line, assuming waves traveling at the speed of light, corresponds to a line 3km long. Needless to say, the magnitude of the problem requires a very low-loss line.

2.1 Attenuation

To demonstrate, let's look at the attenuation of a transmission line. The general expression for attenuation is

$$\alpha = \frac{1}{2} \frac{R_s}{Z_0} \sqrt{\frac{\omega}{2}}$$

where R_s is the skin-effect impedance dependent on the geometry and resistivity of the structure[2].

For a coaxial line the characteristic impedance is

$$Z_0 = \frac{\eta}{2\pi} \ln(b/a)$$

where η is the impedance of free space (377 Ω) and a and b are the inner and outer diameter of the coax respectively. R_s is [2][3]

$$R_s = \frac{\sqrt{j\omega\mu}}{2\pi} \Gamma_r \left(\frac{1}{a} + \frac{1}{b} \right)$$

Γ_r is associated with the skin depth and is

$$\Gamma_r = \begin{cases} 1/\sqrt{j\omega\mu\sigma} & \text{normal conducting metal} \\ 1/\sqrt{j\omega\mu\sigma + 1/\lambda^2} & \text{Type I superconductor} \end{cases}$$

where λ is the penetration depth of the superconductor.

Suppose the line was made of copper ($\sigma = 5.92 \times 10^7 (\Omega \cdot m)^{-1}$) and the inner and outer diameters were 2cm and 3cm respectively. With an 8nsec. rise time, we used 125MHz to be the highest critical frequency, which is about 3 times that of the roll-off frequency. For 125MHz, $\alpha = 7.658 \times 10^{-4}/m$. For a 3km line, this corresponds to a power attenuation of $(1 - e^{-2\alpha \cdot 3000m}) = 98.9\%$.

On the other hand, for the same set of dimensions, a niobium ($\lambda=39\text{nm}^1, \sigma=8.03\text{e}6(\Omega\text{-m})^{-1}$) line yields an α of approximately $5.1025\text{e-}6(1 + j)/\text{m}$. The imaginary term contributes to the inductance of the line. The power attenuation is now 3% at 125MHz which is an upper limit. This is the most compelling reason to use a superconducting line.

3 DESIGN OF A SUPERCONDUCTING LINE

It is clear that attenuation is not an issue with these transmission lines. As it turns out, dispersion becomes the dominant issue, especially at higher frequencies. In addition, a good design should accommodate manufacturing considerations. The size of the transmission line is fundamentally limited by the critical field of the superconducting material. Of the type I conductors, niobium (Nb) is the most generous material since it has the highest critical B -field of .206T¹ at $T=0\text{K}$.

Assuming the distribution for the critical field to be

$$B_c(T) = B_c(0) \left[1 - (T/T_c)^2 \right],$$

the critical field at 4.2K (the usual operating point of liquid helium) is .1635T. The B -field of a structure can be written as $B=gI$ where g is a geometry factor. The condition on g is therefore $g \leq B_c / I$

3.1 Coaxial line

For a concentric coax, $g=\mu/2\pi r$ where r is either the inner or outer radius of the line. This yields the condition that $r > \mu/2\pi B_c$. If one wants to operate at 20% of maximum capacity, then $a=.2\mu/2\pi B_c$ where a is the inner radius. The current on the line can be as high as 1kA. This requires that $a=6.12\text{mm}$.

Since the layer of niobium needs only to be much greater than the penetration depth λ , a thin coat of approximately .5um will be deposited on a copper substrate as shown in Fig. 1. The area between the two conductors will be filled with a material that may or may not be dielectric. The inner conductor should be hollow to allow helium to flow through the tube.

Design 1: The outer conductor radius, for an impedance of 25Ω , is 9.28mm. Design 2: If one puts in a dielectric material of $\epsilon_r=10$, the outer radius goes up to 22.85mm. The advantage of the dielectric is that the length of the line is reduced by a factor of $1/\sqrt{10}$ which yields a length of 948m.

Fig. 1 a) Coaxial line with Nb layers deposited on copper tubes. b) Dimensions of stripline design.

3.2 Stripline

For manufacturing considerations, it may be easier to fabricate striplines instead of coaxial lines. The solution for the stripline problem, both open and closed, has been derived by Primozich[4]. As it turns out, a stripline with a dielectric filler is too dispersive (see section 5). Therefore, the design cannot have a dielectric material. Design 3: The stripline dimensions are shown in the following.

¹ see [2] back cover

3.3 Sheath helix

Sheath helix lines are like coaxial lines, only the inner conductor is wrapped at an angle. These are slow wave structures and ideal for a delay line. However, they are more dispersive than regular coax. Fig. 2 shows two sets of two overlaid 10kV pulses. The solid line is the actual signal at the output of the line and the dotted is the output had there been no dispersion. These plots are derivations based on Lund[5] and are for uncurved lines only.

Design 4: Fig. 2(a) corresponds to a line with $\epsilon_r=1$, $b/a=1.15$, $p(\text{pitch})/a=2.1$, and $v/c=.336$. The line length is reduced to about 1/3 of the original length. Part (b) corresponds to $\epsilon_r=3$, $b/a=1.4$, $p/a=2.75$, and $v/c=.264$. The second line yields a slightly slower wave but at the cost of increased dispersion as evident by the wiggles at the rising part of the pulse.

The sheath helix line is presented here as a backup design in case the stripline is not easy to manufacture because of it's length or the dielectric used for the coaxial line proves to be lossy or difficult to work with.

Fig. 2 a) — 10kV output pulse for sheath helix without dielectric, --- same pulse if line were non-dispersive
b) With dielectric of 10. Wiggles indicate dispersion.

4 JOINT DESIGN

Sections of the line need to be connected with superconducting joints of comparable geometry. An outer and inner sleeve, coated with Nb, is clamped to the outer and inner conductors of the line. The design is shown in fig. 3. The wall thicknesses are uniformly 1mm.

Fig. 3 Design shows ends of inner and outer conductors of coax sections connected with superconducting sleeves.

The 3D finite element frequency domain solver MAXWELL EMINENCE² was used to characterize the joint with 10cm section coax (with no dielectric) connected on both ends. The impedance matrix from the numerical code was used to derive a voltage-current (*ABCD*) matrix which can then be cascaded to model the entire line. In order to achieve an equivalent model for the line as shown in fig.4, the effective matrix is the product of the *ABCD* matrices for both the joints and the sections of line. In other words,

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} A_l & B_l \\ C_l & D_l \end{bmatrix} \begin{bmatrix} A_j & B_j \\ C_j & D_j \end{bmatrix} \begin{bmatrix} A_l & B_l \\ C_l & D_l \end{bmatrix} \dots \begin{bmatrix} A_l & B_l \\ C_l & D_l \end{bmatrix}$$

where the subscript *l* denotes line parameters, and subscript *j* denotes joint parameters. $A=Z_{11}/Z_{12}$, $B=(Z_{11}^2 - Z_{12}^2)/Z_{12}$, $C=1/Z_{12}$, $D=Z_{11}/Z_{12}$.

Fig. 4 a) Block diagram of line and b) it's circuit model.

The impedance matrix of the transmission line portion is simply

$$Z_{11l} = -jZ_0 \cot \gamma l$$

$$Z_{12l} = Z_{11l} / \cos \gamma l$$

The impedance matrix for the joint, as outputted from MAXWELL, leads naturally to equivalent circuit parameters. Z_{12j} is associated with a capacitance and $Z_{11j}-Z_{12j}$ is associated with an inductance. At low frequencies, $Z_{11j}-Z_{12j}$ gives negative results which is unphysical. This is largely a numerical anomaly since it is the result of subtracting two large numbers of comparable values. The impedance values lie outside the error tolerance of the code. Therefore, only the linear part of the curve, between 100-250MHz is used to derive an effective inductance. The two curves yield $L_j = 2.005\text{nH}$ and $C_j = 6.104\text{pF}$. This yields a characteristic impedance $Z_j = \sqrt{L_j/C_j} = 25.63\Omega$, slightly higher than Z_0 , the line impedance. The final circuit is shown in fig. 4b. The impedance parameters are extracted from the *ABCD* matrix.

The output voltage pulses for a line with joints and one without joints are shown for a 3000m line with 9.8m sections and 300 joints. As one can see, the dispersion is barely perceptible. There is a slight shift in the output between the two cases due to the fact that the joints themselves introduce a small extraneous length. (Fig 5)

5 DISPERSION

The dispersion property of conventional coaxial lines and striplines stems from a frequency dependent resistance (skin depth). The previous section has already shown that the effect from skin depth is negligible for a superconducting line. The dispersion on the superconducting transmission line, instead, is caused by wrapping the long line onto either a drum or into a pancake shape. The electromagnetic problem for a curved coax has been solved in [7].

Fig. 5 Comparison of output between line with and without joints.

Fig. 6 a) Phase velocity vs. position in a pancake-wound coax line. b) Its corresponding phase shift at 125MHz.
The propagation constant for a curved coax is as follows

$$\gamma^2 = \gamma_0^2 \left[1 + \frac{A_2}{R^2} + O\left(\frac{1}{R^4}\right) \right]$$

where $\gamma_0 = \omega \sqrt{\mu \epsilon}$, $A_2 = A\omega^2 + B$, and $R = R(s)$. The constants A and B can be found in [7]. For a pancake design, the outer cable radius is, as a function of s , the total length, and R_0 , the inner radius,

$$R(s) = \sqrt{R_0^2 + \frac{2bs}{\pi}}$$

If one wants to find the total dispersion at the end of the line, the propagation constant would need to be integrated over the length of the line since it varies with s . In other words,

$$e^{-j\pi} \rightarrow e^{-j \int_0^s \gamma ds'}$$

Therefore, the phase shift of the signal at the end of the line is

$$\Delta\phi = -\frac{A_2(\omega)}{2} \gamma_0 \int_0^s \frac{ds'}{R^2(s')} = -\gamma_0 \frac{\pi A_2}{4b} \ln \left(1 + \frac{2bs}{\pi R_0^2} \right)$$

The phase velocity and $\Delta\phi$ for design 2 is plotted in Fig. 6. As one can see in Fig. 6b, the phase distortion is less than 10° at the end of the line for a frequency of 125MHz. The striplines were treated as equivalent coax lines to estimate the dispersion due to bending.

The parameters for designs 1-3 were run through a kicker dynamic code to examine the acceptability of the output signal. All 3 designs produced clean output signals. There was a fifth design – a stripline with $\epsilon_r = 10$ – that yielded excessive dispersion.

6 CONCLUSION

Attenuation in long transmission lines made use of superconductors a necessity. The attenuation in the superconducting line is very small, with joints connecting sections of the line contributing to most of it. Dispersion, on the other hand, can arise simply from the geometry of the line. Curved lines and helix lines are subject to dispersion. Two coaxial lines, one stripline, and a sheath helix design were presented. The challenge of fabricating the line will determine which design is ultimately adopted.

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