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STOCHASTIC COOLING RINGS

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2



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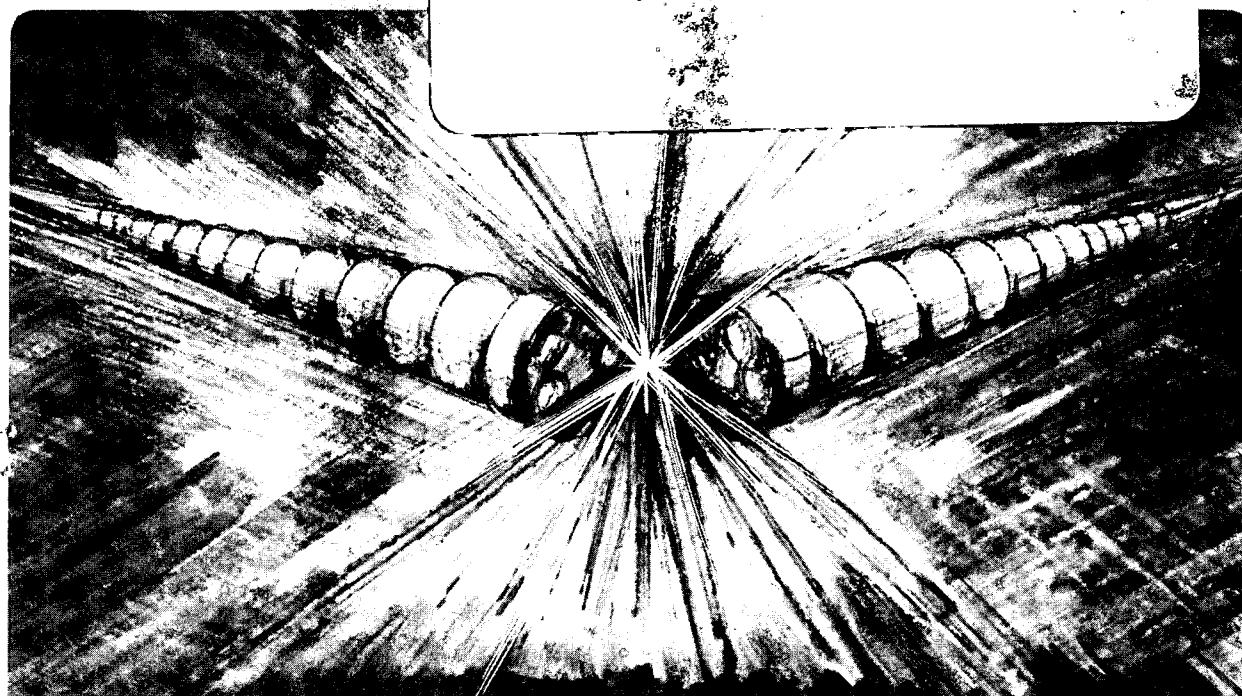
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J.K. Johnson and R. Nemetz

May 1985

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FOR FNAL STOCHASTIC COOLING RINGS*

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Introduction

The anti-proton accumulator and debuncher at FNAL will use stochastic methods to "cool" the beam. Pairs of quarter-wavelength directional-coupler loops [1] are used to detect and kick the beam. The loops are copper plates which are flush with the upper and lower wall of a rectangular beam pipe. The plates, when surrounded by a properly sized pocket, form a 100-ohm transmission-line directional coupler. As the beam passes, a signal which gives position and time information, is induced in the plates. But, because the signal levels are low (<.5 picowatts per pair), a power combiner (usually several primary combiners feeding a secondary combiner) is used to combine the outputs of many loops. Subsequently, the combined signal is amplified, filtered and then fed into a divider, (that is, a combiner operating in reverse). The divider distributes the signal into a different set of loops which modify (kick) the beam's position. Since the loop couplers are arranged linearly, in arrays of various lengths, see Fig. 1, combiners also provide a convenient method of reducing the number of vacuum feedthroughs and preamplifiers and their related costs in performance and dollars. In this note we describe various stripline combiner systems that add the outputs of 4, 8, 16 or 32 loops.

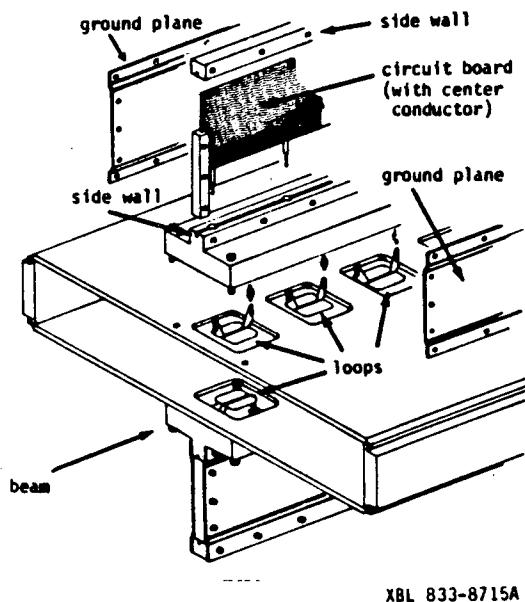


Fig. 1.

The cooling systems operate in one of two frequency bands, 1-2 GHz and 2-4 GHz. The combiners are based on impedance matching theory [2] and the work of Wilkinson [3] and Cohn [4] but modified to meet our particular requirements (see Table 1).

The combiner circuits are made up of the following elements: delay lines for phase matching and coherent signal

addition; bends (most often 45 and 90 degrees) for changes in direction; quarter-wavelength transformers for impedance matching; and combining junctions for adding signals. The components in stripline circuits are situated between a pair of ground planes. This integrated approach has many benefits which will be touched upon as we go on.

Table 1. Combiner Requirements

- operate over an octave bandwidth
- exhibit low loss, smooth amplitude response, and low phase distortion
- incorporate delays to combine individual loop signals in phase
- accommodate loop output port separation of 7 cm, for the 1-2 GHz system, and 4 cm for the 2-4 GHz system
- transform the 100-ohm loop-coupler impedance to 50 ohms
- withstand baking to 150°C
- operate at liquid nitrogen (LN) temperature, 77°K
- be compatible with ultra high vacuum, 10^{-10} torr
- require minimum space inside vacuum envelope

Design

Transmission Medium

We chose the dielectric-supported air strip transmission line (Fig. 2) [5] for its low loss, high phase velocity, the support given wide circuits, and its ease of fabrication. The stripline design allows us to integrate delay lines, combining junctions, and transformers into a single unit within one set of ground planes, thereby eliminating at least two connector pairs and one coaxial cable that would be needed for each loop output were we to combine signals using a single commercial combiner. Reliability should be much higher than with a system of discrete components, virtual (vacuum) leaks from the coax are eliminated, and manufacturing is facilitated since photoetching is accurate and inexpensive. Because the elements of the combiners are simple circuits, we will briefly examine them for some insight into the complete combining circuit.

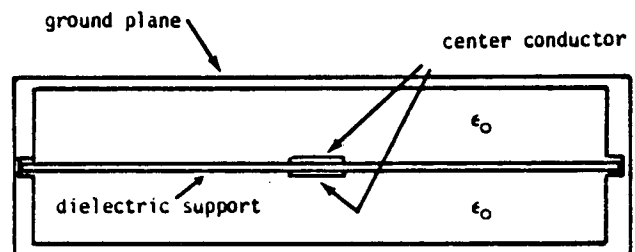


Fig. 2

*This work was supported by the Office of Energy Research, Office of Basic Energy Sciences, U.S. Department of Energy under Contract No. DE-AC03-76SF00098.

Transformers, Combining Junctions, and Bends

We chose the Chebychev or equal-ripple transformer because it has the lowest in-band loss for a given bandwidth [6] and the two-step quarter-wavelength construction as a compromise in size and performance. For a transformation ratio of 100 ohms/50 ohms = 2, the Chebychev impedances are $Z_1 = 61$ ohms and $Z_2 = 82$ ohms.

The combining junctions are the parallel combination (or intersection) of two equal-impedance transmission lines, most often 100 ohms each, which enter the junction at a relative angle of 90 degrees whenever possible. Time Domain Reflectometry (TDR) measurements show that the 90-degree angle gives the least excess capacitance at the junction and, therefore, the least insertion loss (S_{21}).

TDR measurements are also used as a guide for miterring the bends.

Circuit Board Type and Thickness

The circuit board material is a glass-reinforced teflon laminate with two-ounce copper (approximately 2.8 mils) on each side. Its desirable properties include a low dielectric constant, ϵ_r equal to 2.55, for the high phase velocity needed to facilitate in-phase combining of signals, a loss tangent of about .002, for low losses, and a ratio of dielectric to copper thermal expansion of 1.3, necessary to avoid delamination and warpage at extreme temperatures [7]. Its undesirable properties include its high coefficient of thermal expansion compared to the stainless steel of the ground planes, its substantial outgassing rate, and its inadequate stiffness. Outgassing tests show a marginally acceptable rate of about 1.5×10^{-9} torr-liters/sec-cm² after 24 hours of pumping and with no baking. With LN cooling the outgassing rate becomes negligible [8]. We selected a circuit board thickness equal to 20 mils. For our relatively wide boards, a thickness of less than 20 mils allows excessive waviness (deviation from planar).

Ground Plane Spacing

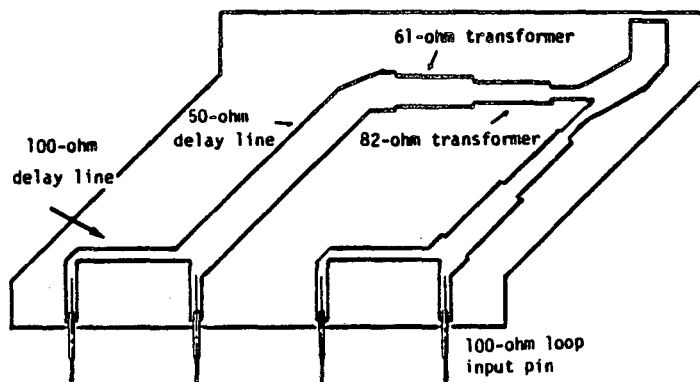
There are several tradeoffs in determining the ground plane spacing (gps), b . For example, a large gps can give a high transmission-line phase velocity making it easier to match the beam phase velocity, (.995c). The ratio of board thickness to gps, controls the phase velocity by determining the fraction of cross section occupied by dielectric. But a large gps can lead to a wide circuit board. Furthermore, to minimize cross-talk between adjacent transmission lines, we found that a separation of at least one gps was required and provides roughly 30 dB isolation.

We chose a gps equal to one centimeter for several reasons. First, it gives a phase velocity of .99c for the 50-ohm transmission line with a circuit board thickness equal to 20 mils and a dielectric constant ϵ_r equal to 2.5. Second, one centimeter is large enough to forgive small deflections of the circuit board from the center of the ground planes thereby obviating a tight tolerance. Third, it is large enough to keep the width of the 100-ohm lines from being too narrow, that is, they are wide enough that the artwork errors (typically ± 7 mils) are less than one percent, and resistive losses are not prohibitive.

Design Example

An example of the design procedure might be instructive. Referring to Fig. 3 we have a board designed to combine (in phase) the signals from four loops. We start by combining the first two (100-ohm) loop outputs at 90 degrees while including the appropriate delay line in the first loop's output. In order that the beam-induced outputs

from the two loops reach the combiner simultaneously (or in phase), the delay line must be as long (electrically) as the travel time of the beam between ports. For the case we are considering, the longitudinal separation of the loop output ports is four centimeters; therefore, with a phase velocity of .977c for the 100-ohm line and a beam beta of .995, the length of the delay line should be 3.928cm. We do the same for the second pair of 100-ohm lines.



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Fig. 3. Center Conductor Pattern for a 4-way Combiner

We now have two 50-ohm lines with a longitudinal separation of eight centimeters. To provide proper phasing of these two signals, we must extend the first 50-ohm line by 2×3.96 cm ($8c = .985c$ for 50-ohm line). Having done this, we transform the two 50-ohm lines to 100-ohms and then combine them. We are now back to 50 ohms and are ready to exit at an appropriate place.

An 8-way combiner simply adds one more set of transformers, a delay line, and another junction. We have built a 16-way combiner, but found that the thermal expansion of the substrate compared with the surrounding array and ground planes causes mechanical problems.

Secondary Combiners

The secondary combiners were required when we found that we were unable to combine the desired number (32) of loop pairs for a given preamp because the width of the resulting board would be too large for the vacuum chamber. The secondary combiners are arranged along side the primaries, to which they are connected by a short length of spline coaxial cable. They differ from the primaries in that they combine 50-ohm inputs and hence the first step is to transform to 100-ohms before the junction. Also, the secondaries use a three-step transformer, permitted by the larger port separation of the inputs, giving less insertion loss over the frequency band than the two-step.

Losses

There are several contributors to the total loss of the combiners. Among them are: conductor loss, dielectric loss, and insertion loss resulting from non ideal elements; transformers, junctions, bends, and connector-board interfaces.

We can estimate the conductor loss by considering our two-part center conductor a single center conductor with a thickness equal to our dielectric's thickness and using Cohn's approximation [9].

Using typical numbers for our 50-ohm 2-ounce combiners at 4 GHz, we get .0015 dB/cm. The conductor's surface finish is neglected in this calculation which will cause the insertion loss estimate to be optimistic.

The dielectric's contribution to the insertion loss is negligible. And the transformers are assumed to be ideal. The S_{11} of the combining junctions (90 degree) has been measured and found to be about 20 dB which gives an insertion loss due to mismatch of .04 dB. The insertion loss of the various bends is perhaps half of this. The insertion loss of the connector-board interface depends heavily on the construction, but is estimated for the trimmed center conductor at .08 dB. Therefore, for the total losses, a_t , in the 2-4 GHz 8-way primary combiner, we have

$$a_t = a_c + a_{\text{junctions}} + a_{\text{bends}} + a_{\text{interface}} \\ = .36 \text{ dB.}$$

Measurements

Our measurements were done on a vector automatic network analyzer (VANA). Typical repeatability is $\pm .05$ dB and uncertainty is $\pm .25$ dB. A full error correction routine is used throughout and all test port connectors are precision 3.5 mm. We use semi-rigid 141 cables for the test port connections taking care not to bend them excessively. One test port of the VANA is connected to the output of the combiner, the other test port is connected to each port in turn with all other ports terminated in 50 ohms. The S_{11} of the terminations is at least 28 dB.

The following parameters are of interest:

- 1) S_{11} at the output port
- 2) group delay of the longest channel
- 3) phase difference between ports
- 4) overall deviation from linear phase
- 5) (mathematical) vector sum of the measured individual combiner branches

For distortionless transmission, flat amplitude response and linear phase are required. Amplitude variations (about the combiners intrinsic power split) are largely the result of multiple reflections from terminations and various mismatches in the combiners at locations such as combining junctions. In commercial combiners isolation resistors, in addition to providing isolation between ports, tend to reduce or smooth the fluctuations by absorbing the useless reflections. Our design contains no isolation resistors because the LN temperatures would cause circuit board stresses that might fracture the resistors. Furthermore, when used in pickups these might prove to be an additional noise source, and when used in kickers they would need heat sinking (because of the high power levels), which is difficult in this particular geometry. The 'vector sum' is a figure of merit wherein at each frequency we take the (mathematical) vector sum of the insertion loss of all (4 or 8) combiner channels corrected for the appropriate beam delay. Table 2 summarizes the results of our measurements on 4-way combiners.

Table 2. Typical Results for Two Combiners*

Frequency range	2-4 GHz	1-2GHz
Return loss [dB]	17	18
Group delay (longest chan.) [cm]	29.5	227.4
Deviation in port separation (equiv. deg. @ fmax)	8	14
Deviation from linear phase (vector sum) [deg]	12	6
S_{21} (vector sum) [dB]	.2	.6

*For more complete results for these and other combiners, see Ref. 10.

Boards In-situ

In use the boards are connected to other non-ideal components that contribute to overall performance. For example, the loops present a S_{11} of about 15 db. The worst S_{11} of the boards alone is 16 db for the 8-way primary at 4 GHz. When terminated by the 15 db loops, the S_{11} at the output of the combiner becomes 10 db.

Another difference is the use of stainless steel in the ground planes of the production arrays versus copper for the test boards. Coaxial cable estimates show that a stainless steel outer conductor at room temperature contributes 65% of the conductor loss. At LN temperatures the stainless steel contribution becomes 95%. It is expected that the conductor losses of our geometry will be similar.

Conclusion

We have described the combiners which will be used in the FNAL anti-proton storage ring. There are many improvements which could be made. Using copper or silver (plated) ground planes is a simple way to substantially reduce losses. Using only 90-degree combining junctions will reduce the insertion loss. Use of Chebychev tapers, in lieu of the stepped transformers, would also reduce insertion loss although at the cost of requiring additional space. Precision connectors having less insertion loss than the SMA connector can be developed. Finally, the center conductor could be made from solid metal plate to avoid the effects of the dielectric. Still, it is clear that the combiners described here will out perform a combination of commercial combiner and coaxial cable delay lines for our application.

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