# HIGH TEMPERATURE SUPERCONDUCTING 8.45-GHz BANDPASS FILTER FOR THE DEEP SPACE NETWORK*

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## ABSTRACT

A high temperature superconducting stripline 8.45 GHz bandpass filter, designed for the Deep Space Network, is presented which has very low insertion loss, indicative of resonators with unloaded Q's greater than 10,000. The performance of this filter shows that practical, high Q multi-resonator devices can be constructed using lanthanum aluminate substrates in a stripline configuration.

## A. Why Superconductivity was Needed

A very low loss, highly selective narrow bandwidth filter is needed for the Deep Space Network to eliminate interfering signals that reduce sensitivity to the required signals. Due to the very weak signals that this network must detect, the input noise figure is minimized by cryogenically cooling the front end below 20K. Thus, high temperature superconducting (HTS) filters are ideal for this application, as very compact designs can be achieved that exhibit minimum insertion losses of fractions of a dB. The filter design needs to minimize not only conductor and dielectric loss, but also radiative loss. Therefore, stripline was chosen as the transmission line medium, with superconducting signal lines and ground planes. This required the packaging of double sided HTS substrates, which was a challenge considering the thermal stresses, the rigid substrates and the very low insertion loss required. The HTS films used in this filter were of the Thallium-Barium-Calcium-Copper-Oxide (TBCCO) variety, deposited on a lanthanum aluminate substrate ($LaAlO_3$) using laser ablation and post deposition thermal processing. The HTS surface resistance of these TBCCO films is less than 0.5 m$\Omega$ at 10 GHz (over 50 times better than OHFC copper) and facilitates the construction of planar HTS resonators with unloaded Q's greater than 10,000 at 10 GHz. HTS planar circuits with these Q's are very useful in constructing narrow bandwidth filters due to the resulting very low minimum insertion loss as shown in Figure 1.

## B. Electrical Design

With regard to the type of transmission-line to be used in the filter, one might be tempted to consider using microstrip lines because of their relative simplicity and their compatibility with planar fabrication techniques. However, microstrip circuits have a major flaw for high-temperature superconductor (HTS) applications in that microstrip circuits always radiate to some degree. In many conventional microstrip circuits the conductor losses are relatively high, and, if the substrate is not too thick, the added loss caused by radiation is not an important consideration. However, in HTS circuits where, often, resonator Q's of the order of 10,000 or more may be desired, the radiation present in microstrip circuits is not acceptable. Thus, in the design of the filter under discussion, stripline was used because it is still compatible with planar fabrication methods and can be entirely enclosed so as to prevent radiation while easily accommodating HTS material in all regions having significant current

After providing a bandwidth safety margin in order to allow for possible errors in dimensions and material parameters, it was decided to design for the band from 8.375 to 8.525 GHz which is centered on 8.45 GHz. It was found that a 5-resonator design with this pass band having a 0.05-dB Chebyshev ripple should more than meet the design requirements. Approximations given in Sec. 4.13 of [1] indicate that if the resonator Q's are 10,000 the added loss due to dissipation should be about 0.16 dB at midband. The loss can be expected to be higher at the band edges. The filter objective was to have a maximum loss of 0.5 dB in the operating band, so with some allowance for loss of perhaps 0.2 dB in the connectors (which are not superconducting) the filter objective is seen to be quite stringent. It is interesting to note that if the resonator Q's were 200, as might be expected for conventional microstrip, the midband loss would be around 8 dB, and, of course, the pass band shape would be entirely rounded out.

At STI we are currently using $LaAlO_3$ substrates with TBCCO HTS transmission lines in our microwave circuits. The $LaAlO_3$ substrates have very low dielectric loss ($\tan\delta \sim 3 \times 10^{-5}$) and a relative dielectric constant of about 24. At 8.45 GHz a half wavelength in this material is only 0.143 inches so a 5-resonator filter, similar to the filter in [2], consisting of capacitively coupled nominally half-wavelength

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resonators will be less than an inch long, and with the narrow bandwidth required, the capacitive gaps will be sufficiently large so as to permit reasonable tolerances. Figure 2 shows the layout of half of the filter. The impedance of the transmission lines within the filter was set at 18.6 ohms in order to give wider lines and better resonator unloaded Q's, and single section quarter-wave transformers were used on the ends to match to 50 ohms. Figure 3 shows a cross-sectional view of the structure which is quite similar to that utilized in [2] Two substrates of $LaAlO_3$ were used with an HTS ground plane on one side of each substrate and the HTS circuit pattern in Figure 2 on the other side. It was necessary to have the center-conductor circuit pattern on both substrates since with the high-dielectric-constant substrates that were used any small air gap between one of the substrates and the center conductor would have caused major disruption in the circuit performance. The substrates where 0.020" thick and 0.104" wide With these dimensions the structure in Figure 3 can begin to propagate the $TE_{10}$ mode at 11.6 GHz. At approximately this frequency the stop band of the filter starts to break down; however, this was not a concern for the present application as the stop band was only needed up to 10 GHz. For filter applications where a broad stop band above the pass band is required other HTS filter techniques are available or mode suppression can be applied.

Design theory for capacitively coupled transmission-line filters of the general type of that in Figure 2 has been known for a long time [1],[3], but to our knowledge the literature does not contain all the information needed for the precise design of such filters which involve sizable coupling gaps. In [2] the series and shunt capacitances shown in Figure 4(a) associated with the gaps were computed using approximate equations due to Oliner (see p. 442 of [1]). However, these equations do not account for the fringing at the outer edges of the gaps. In order to obtain more accurate values for the capacitances in Figure 4(a) we used the program EM [4] which is able to model such cases. To design the filter we used the methods of Secs. 8.02 and 8.03 of [1], which are actually a generalization of the methods used by Cohn in [3]. In [1] it is shown that a bandpass prototype of the form in Figure 5(a) can be designed from a lowpass prototype and the specified fractional bandwidth. In this figure the $J_{k,k+1}$ are admittance inverter parameters and the $b_k$ are resonator slope parameters. The admittance inverters of narrow-band filters present high impedances to the adjacent resonators, and under those conditions the lumped resonators can be replaced by half-wavelength transmission-line resonators which have admittance slope parameters of $b_k = \pi Y_0/2$. The required inverter parameters $J_{k,k+1}$ are easily computed [1], and then the realization of these inverters is accomplished with the aid of Figure 4(b) and the EM program. In Figure 4(b)

$$\phi = -\tan^{-1}\left(2\frac{B_b}{Y_0}+\frac{B_a}{Y_0}\right)-\tan^{-1}\left(\frac{B_a}{Y_0}\right) \qquad (1)$$

and

$$J_{k,k+1} = Y_0\left|\tan\left(\frac{\phi}{2}+\tan^{-1}\frac{B_a}{Y_0}\right)\right| \qquad (2)$$

where $B_a = \omega C_a$ and $B_b = \omega C_b$. Note that the line electrical lengths $\phi$ are negative so they will cancel some of the positive lengths of the adjacent lines In this manner EM and Eqs. (1) and (2) were used until gaps that gave the needed $J_{k,k+1}$ values were found, and then the adjacent lines were shortened by the amount indicated by the associated $\phi$ values.

## C. Packaging and Interface Design

The biggest problem in packaging these HTS circuits is due to the rigid $LaAlO_3$ substrates, with the associated mechanical stresses and electrical grounding problems. Figure 3 shows the assembly concept used. Both substrates had HTS ground planes, each with ohmic gold deposited on the ground plane HTS in order to make contact with the housing. Connection from the upper substrate ground was made to the housing using gap welded gold ribbon, while the gold on the lower substrate ground plane formed a contact directly with the housing ground. The HTS circuit pattern was required on both substrates (due to the problems associated with the air gap) and a pressure contact between the two was maintained using the gasket between the upper substrate and the cover. It proved crucial to maintain pressure on the substrates at all temperatures. If this is not done, then the substrate grounding and connection between the circuit patterns can be degraded resulting in poor electrical performance. The circuit patterns on the two substrates need to be carefully aligned. This was done using cross hair targets etched into the circuit side of the upper substrate, with the negative image etched into the circuit side of the lower substrate. These were located in all four corners of the substrate with windows etched in the ground plane in order to allow for visual alignment. These alignment marks and windows were located in the corners in order to allow enough light through the edges of the substrate so that the cross hair pattern could be seen under the microscope Once visually aligned, the gold ribbon was welded from the upper substrate ground to the housing.

Connection to the filter was by means of a single section quarter-wave microstrip impedance transformed located on the lower substrate circuit pattern. This transformed the 50 $\Omega$ coaxial line impedance to the 18.6 $\Omega$ impedance of the stripline section, resulting in a microstrip line characteristic impedance of $\sqrt{50 \times 18.6} = 30.5\Omega$. The cover over the upper stripline substrate was extended over the microstrip section on to the housing, in order to provide a good ground path for the ground currents. The width of the microstrip line was adjusted to account for the low cover height. Beside performing the impedance transformation, this section of microstrip line was used to make the contact to the coax pin of the connector. The relatively wide microstrip line (0.0185") and the gap between the circuit and the cover (0.020") facilitated the attachment of this ribbon. This was done by welding a gold ribbon from the connector pin on to an ohmic gold pad located on top of the HTS microstrip circuit pattern.

## D. Performance of the Filter

The performance of the filter is shown in Figure 6. The filter was built with one male and one female connector in order to reduce the problems associated with a cryogenic calibration. In order to make the insertion loss calibration, the test connectors were mated directly together and cooled in a liquid nitrogen bath. The test connectors were then directly mated to the filter, thereby providing a measurement of filter performance including connectors. All tests were conducted at 77K by the use of conduction cooling to a liquid nitrogen bath. The measured connector to connector insertion loss at the center frequency was slightly under 0.3 dB, and very flat across the passband (indicative of the high Q resonators). The VSWR was also good, with a worst case return loss of 17 dB. It is clear from the response that the filter is well aligned, due to the symmetry and because all 5 transmission poles can be seen in the return loss response. There is concern about the use of $LaAlO_3$ as the substrate material due to the variations in dielectric constant, and the surface roughness after HTS processing. However, the stripline configuration used showed very good performance, which was thought to be due to the use of 2 substrates to average the dielectric constant and the double circuit pattern to avoid problems due to gaps between the substrates. Several filters were built, all showing good performance, but with some slight variation in performance. This was probably due to small substrate mis-alignment or the effect of variable variable dielectric constant across the substrates.

The performance of the filter from 0.13 GHz to 10 GHz is also shown in Figure 6. This shows an absence of spurious passbands and an out-of-band rejection of at least 80 dB (the limit of the test equipment). The 80 dB rejection bandwidth was measured to be about 1 GHz. As expected, the filter exhibited a degraded stopband starting close to 11 GHz due to the TE10 waveguide mode in the stripline section. The insertion loss of the filter did not measureably vary for input power levels up to 0 dBm. Increasing input power from 0 dBm to +15 dBm resulted in an increased center band loss of about 0.2 dB, with a slightly larger increase in insertion loss at the edges of the passband.


## References

[1] G. L. Matthaei, L. Young, and E. M. T. Jones, Microwave Filters, Impedance-Matching Networks, and Coupling Structures, Artech House, Inc., Norwood, Mass., 1980.

[2] F. J. Winter and J. J. Taub, "High Dielectric Constant Strip Line Band Pass Filters," IEEE Trans. on MTT, vol. 39, pp. 2182-2187, December 1991.

[3] S. B. Cohn, "Direct Coupled Resonator Filters," Proc. IRE, vol. 45, pp. 187-196, February 1957.

[4] EM is a product of Sonnet Software, Inc., Liverpool, NY 13090.

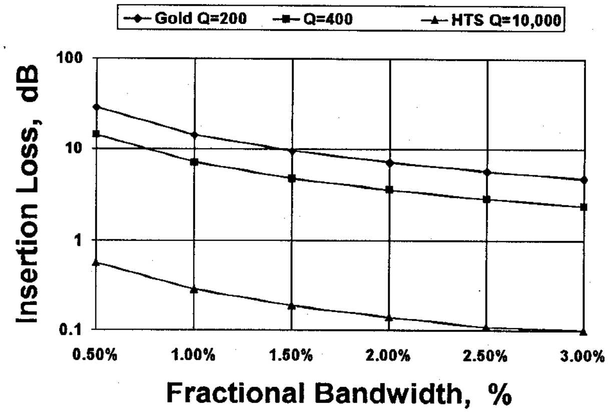




Figure 1. Minimum center band insertion loss of 5th order, 0.05 dB equal ripple Chebyshev bandpass filters.

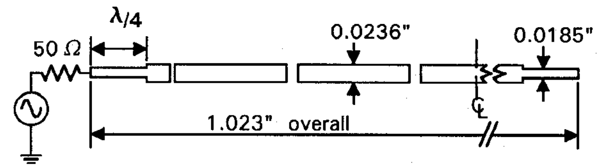



Figure 2. Half of the superconducting five-resonator stripline filter.

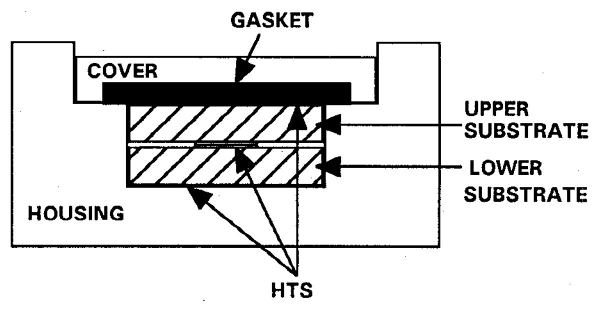
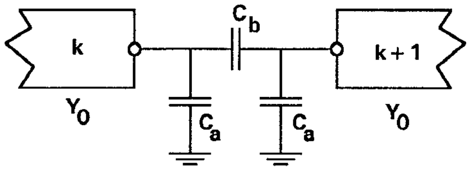
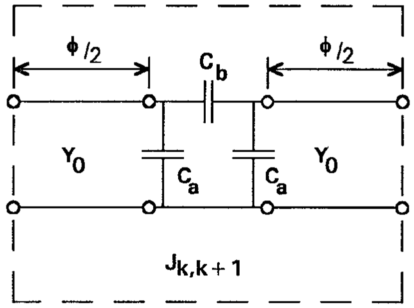



Figure 3. Cross-sectional view of the filter.

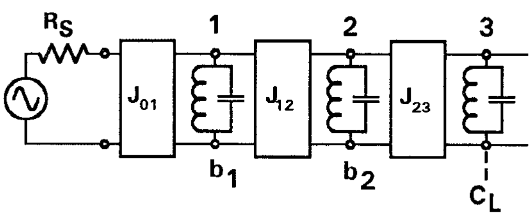



(a)

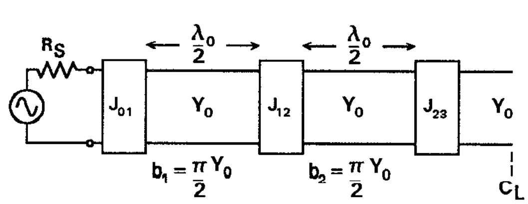



(b)

Figure 4 At (a) is shown a coupling gap with fringing capacitances as referred to the gap edges. At (b) is shown the admittance inverter which is associated with the capacitances.

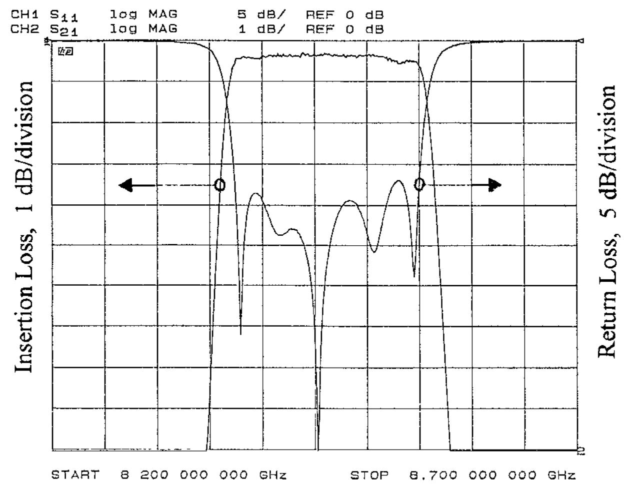



(a)




(b)

Figure 5. The circuit at (a) is a bandpass prototype for the filter in Fig. 1, and at (b) is shown its equivalent with the lumped resonators replaced by transmission lines [1].



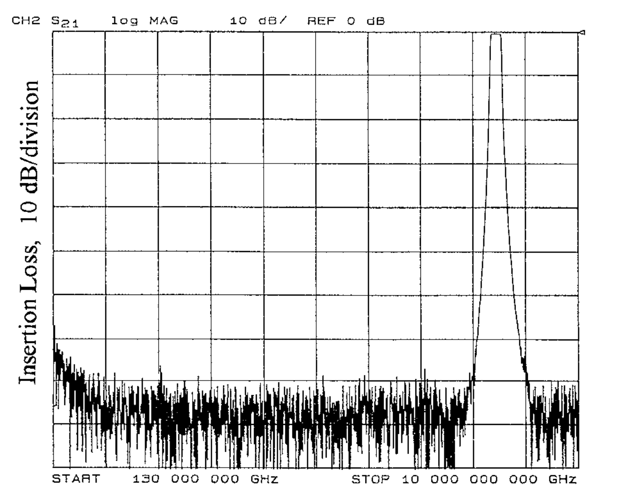




Figure 6. The measured frequency response of the five-resonator filter.