

1                   The Pennsylvania State University  
2                   The Graduate School  
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4   **DEVELOPMENT OF SCALABLE APPROACHES TO NEUTRINO MASS**  
5                   **MEASUREMENT WITH THE PROJECT 8 EXPERIMENT**

6                   A Thesis in  
7                   The Physics Department  
8                   by  
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<sup>20</sup> **Abstract**

<sup>21</sup> Some shit goes here.

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<sup>148</sup> Shout out to all the haters.

<sup>149</sup> **Dedication**

<sup>150</sup> Something heartfelt.

<sup>151</sup> **Chapter 1** |  
<sup>152</sup> **Introduction**

<sup>153</sup> **1.1 Summary**

<sup>154</sup> Neutrinos are one of the fundamental particles that comprise the standard model of  
<sup>155</sup> particle physics and account for a significant fraction of the matter in the universe.  
<sup>156</sup> Neutrinos are the most abundant fermions in the universe, but due to their weak  
<sup>157</sup> interactions neutrinos seldom interact with other particles. Regardless, neutrinos play a  
<sup>158</sup> unique role in the evolution of the early-universe, therefore, a detailed understanding  
<sup>159</sup> of the properties of the neutrino is important to understanding the cosmology of the  
<sup>160</sup> universe as well as understanding the universe at the fundamental particle physics scale.

<sup>161</sup> Unlike other fermions it was unclear that neutrinos had nonzero mass until neutrino  
<sup>162</sup> flavor oscillations were definitively observed in the late 90's and early 00's. Flavor  
<sup>163</sup> oscillations require that neutrinos experience time so that when acted upon by the  
<sup>164</sup> time-evolution operator the initial neutrino state can evolve to a new flavor state. This  
<sup>165</sup> implies that the neutrino flavor states are really a superposition of at least three separate  
<sup>166</sup> neutrino states with well-defined masses. Measurements of neutrino oscillations that have  
<sup>167</sup> taken place over the past couple of decades have measured the differences between  
<sup>168</sup> neutrino mass eigenstates with increasing precision. However, oscillation measurements  
<sup>169</sup> cannot tell us the mass scale of the neutrinos, which is required in order to measure the  
<sup>170</sup> absolute neutrino masses.

<sup>171</sup> The neutrino mass scale remains an unknown quantity in the standard model of  
<sup>172</sup> particle physics. The value of the neutrino mass influences the evolution of the early  
<sup>173</sup> universe and is likely relevant to the energy-scale of new physics responsible for the factor  
<sup>174</sup> of  $10^{-6}$  difference between the neutrino and electron masses. A model-independent way  
<sup>175</sup> to measure the neutrino mass is to measure the tritium beta-decay spectrum near its  
<sup>176</sup> endpoint. Energy conservation requires that the neutrino mass carry away some kinetic  
<sup>177</sup> energy from the beta-decay electron in the form of its mass, which causes a distortion in

178 the shape of the tritium beta-decay spectrum near the endpoint. The isotope tritium has  
179 many advantages for this measurement, and has been used by the KATRIN collaboration  
180 to perform the most sensitive direct neutrino mass measurement to date.

181 KATRIN represents the state-of-the-art experiment in the current generation of  
182 neutrino mass direct measurement experiments and has a final projected sensitivity to  
183 neutrino masses  $m_\nu < 200$  meV. This sensitivity does not fully exhaust the allowed  
184 parameter space of neutrino masses under the normal and inverted neutrino mass  
185 ordering scenarios, which motivates the development of a next generation of neutrino  
186 mass measurement experiments.

187 The Project 8 collaboration is developing a next-generation neutrino mass direct  
188 measurement experiment designed to be sensitive to  $m_\nu < 40$  meV. This sensitivity  
189 is sufficient to exhaust the range of neutrino masses allowed under the inverted mass  
190 ordering regime. Project 8 intends to achieve its sensitivity goal utilizing two technologies  
191 that are novel to the space of direct neutrino mass measurement — atomic tritium and  
192 cyclotron radiation emission spectroscopy (CRES). Atomic tritium is required in order to  
193 avoid systematic broadening the tritium beta-decay spectrum caused by the final state  
194 of the  ${}^3\text{He}^+ \text{-T}$  molecule, and the CRES technique enables a differential measurement of  
195 the tritium spectrum that is background-free and able to be directly integrated with the  
196 atomic tritium source.

197 The Project 8 collaboration is currently engaged in a research and development  
198 program intended to simultaneously develop the atomic tritium and CRES technologies  
199 so that they can be combined in a next-generation experiment. This past year (2022)  
200 Project 8 has used the CRES technique to measure the molecular tritium beta-decay  
201 spectrum and place an upper limit on the neutrino mass:  $m_\beta \leq 152$  eV. This measurement,  
202 while not competitive scientifically, represents the first proof-of-principle that the CRES  
203 technique can be used to measure the neutrino mass.

204 The future goals of the Project 8 collaboration are to develop the technologies  
205 and techniques necessary to scale-up the volumes in which CRES measurements can  
206 be performed. Project 8's first neutrino mass measurement with CRES utilized a  
207 measurement volume on the cubic-centimeter scale, however, sensitivity calculations  
208 estimate that an experiment sensitive to neutrino masses of 40 meV will require several  
209 tens of cubic-meters of experiment volume filled with atomic tritium. Developing a new  
210 approach to performing CRES measurements that can be successfully scaled to these  
211 volumes is a necessary step towards Project 8's neutrino mass measurement goal, and is  
212 the primary topic of my dissertation research.

213 A parallel development is the technology necessary to produce, cool, trap, and  
214 recirculate a supply of atomic tritium that is compatible with CRES measurements. The  
215 atomic tritium system is equally important as the large-volume CRES measurement  
216 technology, but it will not be the focus of this dissertation since I did not contribute  
217 significantly to this effort.

218 The Project 8 collaboration has identified two scalable approaches to neutrino mass  
219 measurement using the CRES technique. One approach is to use an array of antennas  
220 that surrounds a volume of trapped atomic tritium that can perform CRES measurements  
221 by collection the cyclotron radiation emitted by beta-decay electrons into free-space. The  
222 other approach uses a resonant cavity filled with atomic tritium to perform CRES by  
223 measuring the excitation of resonant cavity modes caused by the motion of electrons  
224 trapped inside the cavity volume.

225 The cavity and antenna approaches to CRES have been studied in detail over the past  
226 five years, and, while both approaches offer a physically viable path towards a 40 meV  
227 neutrino mass measurement the collaboration has elected to pursue the cavity approach  
228 for the foreseeable future. The major advantage of the cavity approach is a significant  
229 reduction in the cost and complexity of the experiment design and data analysis, which  
230 provides a less risky path towards Project 8’s scientific goals.

231 In this dissertation I summarize my most impactful contributions to the research and  
232 development of antenna array and cavity CRES. In short these contributions are

- 233 • the development and analysis of signal reconstruction algorithms for antenna array  
234 CRES, which provided key inputs to sensitivity analyses of antenna array CRES  
235 experiments,
- 236 • the development of a specialized antenna designed to synthesize fake CRES radia-  
237 tion, which enabled bench-top testing and validation of the antenna array CRES  
238 technique,
- 239 • the development of an open-cavity design for CRES measurement whose mode  
240 structure can be tuned using perturbations that modify the impedance of the cavity  
241 walls. The development of this cavity concept was one of many developments that  
242 eventually lead to the adoption of cavities as the CRES technology of choice for  
243 the future of Project 8.

## <sup>244</sup> 1.2 Outline

<sup>245</sup> The outline of this dissertation is as follows. In Chapter 2 I provide an introduction to  
<sup>246</sup> the basic physics of neutrinos and beta-decay, which provides context for a discussion of  
<sup>247</sup> various methods to measure the neutrino absolute mass scale.

<sup>248</sup> Chapter 3 is an overview of the CRES technique and the Project 8 collaboration.  
<sup>249</sup> I highlight the Project 8 Phase II experiment, which was the first measurement of  
<sup>250</sup> the tritium beta-decay spectrum with CRES, and I discuss the planned research and  
<sup>251</sup> development for an antenna array CRES experiment in Phase III of the Project 8  
<sup>252</sup> collaboration’s experiment plan. I end Chapter 3 with a discussion of the pilot-scale and  
<sup>253</sup> Phase IV experiments, that will combine a scalable CRES measurement technology with  
<sup>254</sup> atomic tritium and measure the neutrino mass with 40 meV sensitivity.

<sup>255</sup> Chapter 4 discusses the first of the contributions mentioned above, which is the  
<sup>256</sup> development of signal reconstruction techniques for antenna array CRES and an antenna  
<sup>257</sup> array demonstrator experiment called the FSCD. I discuss the important tools that Project  
<sup>258</sup> 8 uses to simulate antenna array CRES before introducing three signal reconstruction  
<sup>259</sup> algorithms that can be used to detect CRES signals using the array. I end Chapter 4  
<sup>260</sup> with a paper that summarizes a detailed analysis and comparison of the signal detection  
<sup>261</sup> performance of each algorithm.

<sup>262</sup> Chapter 5 describes my contributions to the development of antennas and an antenna  
<sup>263</sup> measurement system for Project 8, which is the second major contribution of this  
<sup>264</sup> dissertation. I begin with a general overview of basic principle of antennas and antenna  
<sup>265</sup> measurements, before including a paper that describes the development of unique antenna  
<sup>266</sup> designed to mimic the cyclotron radiation emitted by electrons in free-space when trapped  
<sup>267</sup> in a magnetic field. I call this antenna the synthetic cyclotron radiation antenna (SYNCA)  
<sup>268</sup> and its main purpose is to serve a fake electron for laboratory validation measurements  
<sup>269</sup> of Project 8’s antenna array CRES simulations. Chapter 5 ends with an overview  
<sup>270</sup> of laboratory measurements of a prototype antenna array that were compared with  
<sup>271</sup> simulations to provide upper bounds on reconstruction errors caused by imperfections in  
<sup>272</sup> real-life measurements.

<sup>273</sup> Chapter 6 discusses the cavity approach to CRES, which was adopted as the preferred  
<sup>274</sup> CRES technology for Phase IV late into my dissertation work. The chapter stars by  
<sup>275</sup> discussing resonant cavities in general before introducing the operating principles of the  
<sup>276</sup> cavity approach to CRES. I end the chapter by discussing a study of and open-cavity  
<sup>277</sup> design that could be used for CRES measurements and integrated with atomic tritium

<sup>278</sup> and an electron gun calibration source for the pilot-scale and Phase IV experiments.

<sup>279</sup> Finally, in Chapter 7 I conclude by briefly discussing the future directions of the  
<sup>280</sup> Project 8 collaboration as we continue towards a direct measurement of the neutrino  
<sup>281</sup> mass.

<sup>282</sup> **Chapter 2 |**

<sup>283</sup> **Neutrinos and Neutrino Masses**

<sup>284</sup> **2.1 Introduction**

<sup>285</sup> In this chapter I provide a cursory overview of background information relevant to  
<sup>286</sup> neutrinos and neutrino mass measurements.

<sup>287</sup> In Section 2.2 I provide some background information on the history of neutrinos and  
<sup>288</sup> beta-decay. In Section 2.3 I describe the discover of neutrino oscillations, which proved  
<sup>289</sup> unambiguously that neutrinos have non-zero masses. In Section 2.4 I discuss the current  
<sup>290</sup> state of the theoretical understanding of neutrino masses in the standard model. Lastly,  
<sup>291</sup> in Section 2.5 I discuss methods for measuring the absolute scale of the neutrino mass.

<sup>292</sup> **2.2 Neutrinos and Beta-decay**

<sup>293</sup> Late in the 19th century the phenomena of radioactivity was first observed in experiments  
<sup>294</sup> performed by Henri Becquerel with uranium, and further studied using thorium and  
<sup>295</sup> radium by Marie and Pierre Curie. Early work in radioactivity classified different forms  
<sup>296</sup> of radiation based on it's ability to penetrate different materials. Rutherford was the first  
<sup>297</sup> to separate radioactive emissions into two types, alpha and beta radiation. Alpha rays  
<sup>298</sup> can be easily stopped by a piece of paper or thin foil of metal, whereas beta radiation  
<sup>299</sup> could penetrate metals several millimeters thick. Later a third form of radiation was  
<sup>300</sup> identified by Villard, which was still more penetrating, and was eventually termed gamma  
<sup>301</sup> radiation by Rutherford.

<sup>302</sup> When these forms of radioactivity were first discovered it was unclear what physically  
<sup>303</sup> constituted an alpha, beta, or gamma particle. Experiments with radioactivity in  
<sup>304</sup> magnetic fields was eventually able to identify the charge composition of different forms  
<sup>305</sup> of radiation. In particular, experiments by Becquerel identified that beta radiation had

306 an identical charge-to-mass ratio to the electron discovered by Thompson in his work on  
307 cathode rays. This was strongly suggestive that beta particles were indeed electrons.

308 Further studies of beta radiation lead to the discovery that radioactivity resulted  
309 in the transmutation of elements caused by the decay of a heavier nucleus to a lighter  
310 species. One feature of beta radiation, which we now properly call beta-decay, that  
311 was different from alpha-decays and gamma radiation is that the electrons produced by  
312 beta-decay have a continuous spectrum of kinetic energies, whereas, alpha and gamma  
313 particles are emitted with discrete energies. This feature of beta-decay was first observed  
314 by Chadwick in 1914, and was extremely puzzling at the time since the continuous  
315 spectrum apparently violates energy conservation.

316 Famously, in 1930 Pauli proposed the existence of a new neutral particle, which he  
317 termed the "neutron", that was also produced during beta-decay in order to resolve the  
318 missing energy problem posed by the beta-decay spectrum. Because this particle carried  
319 no charge, it was hypothesized at the time that it had simply not been observed in any  
320 experiments up to that time. This "neutron", which was initially estimated to have a  
321 mass no larger than that of an electron, was eventually renamed the "neutrino" by Fermi  
322 after the discovery of the neutron by Chadwick in 1932. Later, in 1933, Fermi developed  
323 a quantum mechanical theory for beta-decay in which both an electron and neutrino are  
324 produced by the decay of a neutron to a proton inside the radioactive nucleus.

325 Little more than a speculation when first introduced, indirect evidence for the existence  
326 of neutrinos was obtained in 1938 by the simultaneous observation of the electron and  
327 recoiling nucleus in cloud chambers by Crane and Halpern. However, it wasn't until the  
328 Cowan-Reines experiment in 1956 that direct evidence for the existence of neutrinos  
329 was observed by detecting the inverse beta-decays caused by neutrinos from a nuclear  
330 reactor interacting with protons contained in water molecules. The difficulty in detecting  
331 neutrinos is caused by their weak interactions with other particles. Further, experiments  
332 revealed that different types of neutrinos existed based on the nature of the leptons  
333 produced in neutrino charged-current interactions, but the existence of a neutrino mass  
334 remained an open question that would take more than 40 year to resolve.

## 335 **2.3 Neutrino Oscillations**

336 The first hint of neutrino flavor transitions or neutrino oscillations was indicated by  
337 the solar neutrino problem, which referred to discrepancies between the predicted flux  
338 of  $\nu_e$  from the standard solar model and measurements of the solar neutrino flux such

339 as the famous experiment at the Homestake mine by Ray Davis Jr. and collaborators  
 340 in the 1960's. Essentially, fewer electron-type neutrinos than expected were being  
 341 observed from the sun. Finally, in the early 2000's the SNO experiment was able to  
 342 resolve the solar neutrino problem by identifying neutrino oscillations as the cause of  
 343 the observed deficit. Furthermore, measurements of the atmospheric flux of neutrinos by  
 344 the Super-Kamiokande experiment and others revealed that fewer muon-type neutrinos  
 345 survived passage through the earth than expected providing strong evidence for neutrino  
 346 oscillations for both flavors.

347 The origin of neutrino oscillations is that the weak eigenstates are distinct from the  
 348 mass eigenstates. The neutrino mass eigenstates represent physical particles in the sense  
 349 that they are solutions to the free-particle Hamiltonian, whereas, the neutrino weak  
 350 eigenstates correspond to the neutrino states that interact via the weak charged-current  
 351 interaction. The neutrino weak eigenstates are a linear superposition of the neutrino  
 352 mass eigenstates

$$\nu_\ell = \sum_i U_{\ell i} \nu_i, \quad (2.1)$$

353 where  $\ell = e, \mu, \tau$  and  $i = 1, 2, 3$ . The matrix elements  $U_{\ell i}$  are the elements of the  
 354 Pontecorvo-Maki-Nakagawa-Sakata (PMNS) matrix that describes the mixing between  
 355 the neutrino flavor and mass states.

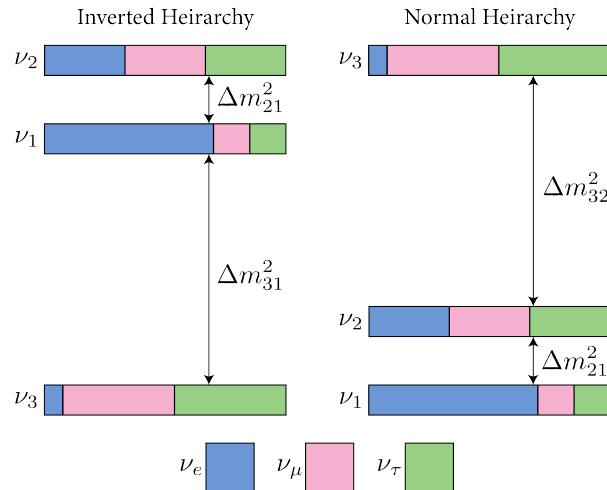


Figure 2.1: A diagram of two different neutrino mass ordering scenarios. In the inverted hierarchy (inverted mass ordering) the lightest neutrino mass is  $m_3$ , whereas, in the normal hierarchy (normal mass ordering)  $m_1$  is the lightest neutrino. What cannot be measured by neutrino oscillations is the neutrino absolute mass scale, which is essentially the mass of the lightest neutrino mass eigenstate.

356 One standard parameterization of the PMNS matrix is

$$\begin{aligned}
U_{PMNS} &= \begin{bmatrix} U_{e1} & U_{e2} & U_{e3} \\ U_{\mu 1} & U_{\mu 2} & U_{\mu 3} \\ U_{\tau 1} & U_{\tau 2} & U_{\tau 3} \end{bmatrix} \\
&= \begin{bmatrix} 1 & 0 & 0 \\ 0 & c_{23} & s_{23} \\ 0 & -s_{23} & c_{23} \end{bmatrix} \begin{bmatrix} c_{13} & 0 & s_{13}e^{-i\delta} \\ 0 & 1 & 0 \\ -s_{13}e^{i\delta} & 0 & c_{13} \end{bmatrix} \begin{bmatrix} c_{12} & s_{12} & 0 \\ -s_{12} & c_{12} & 0 \\ 0 & 0 & 1 \end{bmatrix} \\
&\quad \times \begin{bmatrix} e^{i\alpha_1/2} & 0 & 0 \\ 0 & e^{i\alpha_2/2} & 0 \\ 0 & 0 & 1 \end{bmatrix}, \tag{2.2}
\end{aligned}$$

357 where  $c_{ij} = \cos \theta_{ij}$  and  $s_{ij} = \sin \theta_{ij}$ . The parameters  $\alpha_1$  and  $\alpha_2$  are only included in the  
358 PNMS matrix if neutrinos are Majorana particles, something which represents a current  
359 area of research in neutrino physics. The phase  $\delta$  quantifies the degree of CP-violation  
360 in the neutrino sector. Including the Majorana phases the PMNS matrix contains six  
361 independent parameters. In addition, neutrino oscillation probabilities depend on the  
362 squared mass differences between neutrino mass eigenstates

$$\Delta m_{ij}^2 = m_i^2 - m_j^2, \tag{2.3}$$

363 where  $ij = 12, 32, 31$  respectively. Because  $\Delta m_{32}^2 = \Delta m_{31}^2 - \Delta m_{21}^2$ , this adds an additional  
364 two parameters that must be constrained by neutrino oscillations.

365 A giant experimental effort over the past couple of decades has greatly contained the  
366 majority of parameters in the PMNS matrix, many to relative uncertainties of only a  
367 few percent. However, some parameters still remain relatively unconstrained, which is  
368 the origin of the current uncertainty in the ordering of the neutrino masses (see Figure  
369 2.1). The neutrino masses can be organized by their relative mass. The current neutrino  
370 oscillation data can confirm that  $m_2 > m_1$ , however, the sign of  $\Delta m_{32}^2$  is still unknown.  
371 This leads to two scenarios where neutrino masses follow the ordering  $m_3 > m_2 > m_1$ ,  
372 which is called the normal mass ordering (NMO), or alternatively neutrino masses may  
373 be ordered  $m_2 > m_1 > m_3$ , which is called the inverted mass ordering (IMO). Next-  
374 generation neutrino oscillation experiments such as JUNO, Hyper-Kamiokande, and  
375 DUNE are poised to resolve this ambiguity in the coming years.

376 Neutrino oscillation probabilities are only sensitive to the neutrino masses via the  
377 squared mass differences. Therefore oscillation probabilities are unaffected by the absolute

scale of the neutrino mass. However, oscillations can be used to obtain a lower bound on the neutrino masses by setting the mass of the lightest neutrino mass state to zero. This results in different lower limits depending on the ordering of the neutrino mass states. Current best-fit values with  $1\sigma$ -uncertainties for the squared mass differences are

$$\Delta m_{21}^2 = (7.42^{+0.21}_{-0.20}) \times 10^{-5} \text{ eV}^2, \quad (2.4)$$

$$\Delta m_{31}^2 = (2.5176^{+0.026}_{-0.028}) \times 10^{-3} \text{ eV}^2 \text{ (NMO)}, \quad (2.5)$$

for the normal mass ordering, and in the case of the inverted ordering we have

$$\Delta m_{32}^2 = (-2.498^{+0.028}_{-0.028}) \times 10^{-3} \text{ eV}^2 \text{ (IMO).} \quad (2.6)$$

By letting the lightest neutrino mass in each ordering scenario ( $m_{\text{least}}$ ) take on a range of values one can visualize the relative masses of the neutrinos as a function of  $m_{\text{least}}$  (see Figure 2.2).

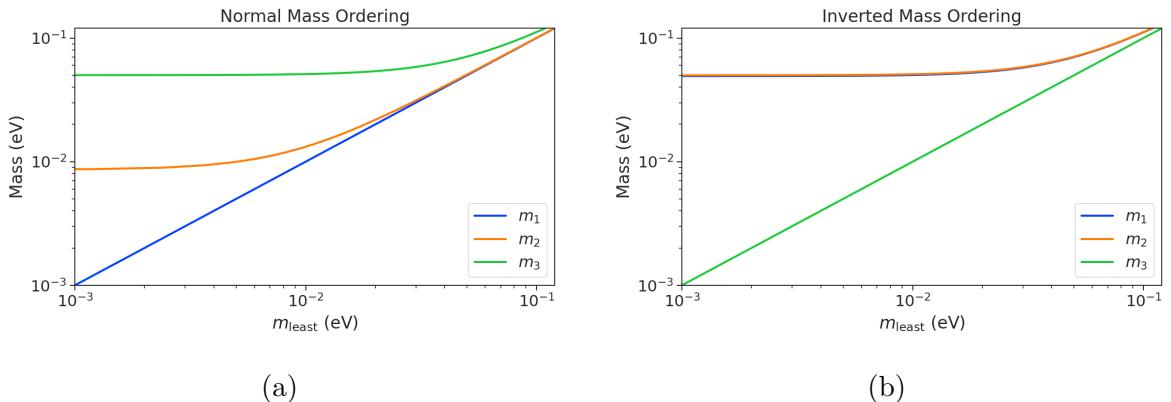


Figure 2.2: The masses of the neutrinos as a function of the lightest neutrino mass in both the normal (a) and inverted (b) mass ordering regimes.

## 2.4 Neutrino Masses in the Standard Model

Neutrinos are spin 1/2 particles and in modern quantum field theory spin-1/2 particles, or fermions, are described using the Dirac equation.

$$(i\hbar\gamma^\mu\partial_\mu - mc)\psi(x) = 0, \quad (2.7)$$

389 where the field that describes the particle is denoted as  $\psi(x)$ . In the standard fermions ac-  
 390 quire mass through the Yukawa interaction, which add to the standard model Lagrangian  
 391 terms of the form

$$\mathcal{L}_{\text{Yukawa}} = -Y_{ij}^\ell \bar{L}_{Li} \phi E_{Rj} + \text{h.c.}, \quad (2.8)$$

392 where  $Y_{ij}^\ell$  is an element of the  $3 \times 3$  Yukawa coupling matrix for leptons,  $L_{Li}$  is the  
 393 left-handed lepton doublet for generation  $i$ ,  $\phi$  is the Higgs doublet, and  $E_{Rj}$  is the  
 394 right-handed lepton field for generation  $j$ . In the standard model neutrinos are only  
 395 represented as left-handed neutrinos and right-handed antineutrinos, consistent with  
 396 experimental observations. Since there are no right-handed neutrino singlet fields and  
 397 no Yukawa interaction terms for neutrinos are strictly massless, and non-zero neutrino  
 398 masses is evidence for physics beyond the standard model. For the charged leptons, the  
 399 Yukawa interaction leads to masses of the form

$$m_{ij}^\ell = Y_{ij}^\ell \frac{v}{\sqrt{2}}, \quad (2.9)$$

400 where  $v$  is the Higgs vacuum expectation value.

401 The observation of massive neutrinos motivates the extension of the standard model  
 402 to explain the origin of neutrino masses, which can be approached in different way, but  
 403 all methods add additional degrees of freedom to the standard model. One approach  
 404 is to introduce to the standard model a right-handed neutrino field that allows one to  
 405 introduce Yukawa terms of the form

$$\mathcal{L}_{\nu \text{Yukawa}} = -Y_{ij}^\ell \bar{L}_{Li} \phi \nu_{Rj} + \text{h.c.} \quad (2.10)$$

406 where  $\nu_{Rj}$  is the right-handed neutrino singlet. Because experimental evidence strongly  
 407 predicts only three active neutrinos these additional neutrinos are sterile and do not in-  
 408 teract via the strong, weak, or electromagnetic interactions. After spontaneous symmetry  
 409 breaking, the Yukawa interaction leads to mass terms given by

$$\mathcal{L}_D = -M_{Dij} \bar{\nu}_{Ri} \nu_{Lj} + \text{h.c.}, \quad (2.11)$$

410 which is called a Dirac mass term. One of the issues with constructing neutrino masses  
 411 in this way is that the required Yukawa couplings are at least a factor of  $10^6$  smaller than  
 412 that of an electron, which begs the question: why are the Yukawa couplings so small for  
 413 the neutrinos?

414 An alternative approach is to allow the neutrinos to have a Majorana mass, which is

<sup>415</sup> possible because neutrinos are electrically neutral particles. The Majorana mass terms  
<sup>416</sup> for the neutrino have the form

$$\mathcal{L}_M = -\frac{1}{2}(M_{Rij}\bar{\nu}_{Ri}\nu_{Rj}^c M_{Lij}\bar{\nu}_{Li}\nu_{Lj}^c) + \text{h.c.}, \quad (2.12)$$

<sup>417</sup> where  $M_{Rij}$  and  $M_{Lij}$  are right-handed and left-handed Majorana mass matrices. A  
<sup>418</sup> consequence of neutrinos being Majorana particles is lepton number violation, which  
<sup>419</sup> predicts the occurrence of neutrino-less double beta-decay at a rate proportional to the  
<sup>420</sup> neutrino mass.

<sup>421</sup> In the most general case neutrinos have both Dirac and Majorana mass terms, which  
<sup>422</sup> allows one to generate neutrino masses with Yukawa couplings similar to the rest of  
<sup>423</sup> the standard model. Considering just one generation of neutrinos for illustration, the  
<sup>424</sup> combined Lagrangian can be written as

$$\mathcal{L}_{D+M} = -m_D\bar{\nu}_R\nu_L - \frac{1}{2}(m_L\bar{\nu}_L\nu_L^c + m_R\bar{\nu}_R\nu_R^c) + \text{h.c.}, \quad (2.13)$$

<sup>425</sup> or equivalently,

$$\mathcal{L}_{D+M} = -\frac{1}{2} \begin{bmatrix} \bar{\nu}_L & \bar{\nu}_R^c \end{bmatrix} \begin{bmatrix} m_L & m_D \\ m_D & m_R \end{bmatrix} \begin{bmatrix} \nu_L^c \\ \nu_R \end{bmatrix} + \text{h.c..} \quad (2.14)$$

<sup>426</sup> An example mass generation mechanism with this approach is the Type-I see-saw  
<sup>427</sup> mechanism, in which we take  $m_L = 0$  and  $m_R \gg m_D$ . By diagonalizing Equation 2.14  
<sup>428</sup> one obtains the mass eigenvalues that represent the physical masses of the neutrinos.  
<sup>429</sup> The light neutrino mass eigenstate, which represents the observed neutrino mass, has a  
<sup>430</sup> mass given by

$$m_1 \approx \frac{m_D^2}{m_R}, \quad (2.15)$$

<sup>431</sup> and the heavy neutrino mass eigenstate, which represents the unobserved sterile neutrino,  
<sup>432</sup> has a mass

$$m_2 \approx m_R. \quad (2.16)$$

<sup>433</sup> For  $m_D$  similar to the other quark or lepton masses, one obtains physical neutrino masses  
<sup>434</sup> consistent with observations from sterile neutrino masses of  $m_R \approx O(10^{15})$  GeV. This  
<sup>435</sup> mass scale is well beyond the capabilities of modern particle accelerators.

## <sup>436</sup> 2.5 Neutrino Absolute Mass Scale

<sup>437</sup> The neutrino absolute mass scale or simply "neutrino mass" cannot be probed with  
<sup>438</sup> neutrino oscillations, since oscillation probabilities are determined by the squared mass  
<sup>439</sup> differences between neutrino mass eigenstates, therefore, alternative techniques are needed  
<sup>440</sup> to perform an effective measurement of the neutrino mass.

### <sup>441</sup> 2.5.1 Limits from Cosmology

<sup>442</sup> In the  $\Lambda$ CDM model, which summarizes our current cosmological understanding of our  
<sup>443</sup> universe, the mass-energy content of the universe is composed of approximately 27%  
<sup>444</sup> dark matter and only 5% normal matter including neutrinos. From this observation, a  
<sup>445</sup> rough limit on the neutrino mass can be obtained from the condition that neutrinos are  
<sup>446</sup> not responsible for the entirety of the matter content of the universe. Using only this  
<sup>447</sup> condition one can constrain the neutrino mass to be ...

<sup>448</sup> A prediction of the  $\Lambda$ CDM model is that the universe originated from a single  
<sup>449</sup> expansion event colloquially called the "Big Bang". In the Big Bang scenario, our  
<sup>450</sup> universe originated as a hot spacetime singularity, which abruptly experience rapid  
<sup>451</sup> expansion in a process called inflation. After the inflationary epoch the universe entered  
<sup>452</sup> the reheating phase where the potential energy responsible for inflation decays into  
<sup>453</sup> standard model particles such as electrons, quarks, and gluons. The universe continued to  
<sup>454</sup> expand in size resulting in a decrease in energy density and lower temperature. Eventually  
<sup>455</sup> the temperature of the universe decreased enough to allow the formation of protons,  
<sup>456</sup> neutrons, and other baryons from quarks and gluons produced from the decays of the  
<sup>457</sup> inflationary fields.

<sup>458</sup> Also produced during the Big Bang are electrons, neutrinos and other leptons as  
<sup>459</sup> well as a population of photons. These particles are kept in thermal equilibrium with  
<sup>460</sup> the rest of the quark-gluon plasma through interactions that take place at the high  
<sup>461</sup> temperatures and densities of the early universe. However, as the universe continues  
<sup>462</sup> to expand it's density and temperatures decreases leading to the eventual decoupling  
<sup>463</sup> of photons and leptons from the quarks and gluons. A prediction of inflation is that  
<sup>464</sup> this population of photons produced during the Big Bang should still be present, but  
<sup>465</sup> with a significantly reduced temperature due to the expansion of the universe. This is  
<sup>466</sup> consistent with the observation of the CMB (cosmic microwave background), which is a  
<sup>467</sup> population of microwave radiation with a blackbody temperature of 2.7 K. The CMB  
<sup>468</sup> is extremely uniform in all directions with slight anisotropies that can be analyzed to

study the evolution of the early universe. A series of experiments have measured the CMB with increasing levels of precision, which has lead to a significant increase in our current understanding of cosmology.

In addition to the CMB, inflation predicts the existence of a  $C\nu B$  (cosmic neutrino background), which are the remnant neutrinos produced during the Big Bang. Since neutrinos only interact via the weak force, they decouple from the hot Big Bang plasma at an earlier time than the CMB radiation. The temperature at which the  $C\nu B$  decouples depends on the neutrino rest mass. Neutrinos play a somewhat unique role in the  $\Lambda$ CDM model due to the fact that neutrinos act as radiation early in the universe but as matter in the late universe. This leads to unique signatures that impact anisotropies of the CMB as well as the distribution of matter in the universe. By combining measurements of the CMB with measurements of the large-scale structure (LSS) of the universe one can constrain the neutrino mass scale by fitting these datasets with the  $\Lambda$ CDM model. This analysis results in some of the most stringent constraints on the neutrino mass. A recent analysis was able to constrain the neutrino mass scale to

$$\Sigma_{m_\nu} \equiv \sum_i m_i < 0.12 \text{ eV}, \quad (2.17)$$

where  $m_i$  are the neutrino mass eigenstates.

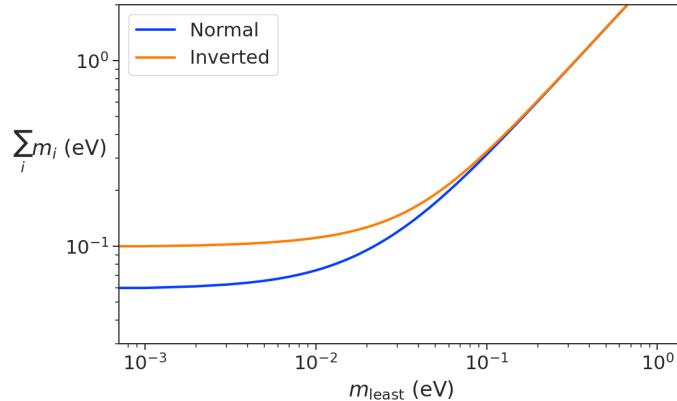


Figure 2.3: The neutrino mass observable measured by cosmology as a function of the lightest neutrino mass eigenstate.

The observable  $\Sigma_{m_\nu}$  constrains the neutrino mass by setting the mass of the lightest neutrino mass eigenstate ( $m_{\text{least}}$ ). In the normal mass ordering  $\Sigma_{m_\nu}$  can be rewritten in

<sup>487</sup> the form

$$\Sigma_{m_\nu} = m_{\text{least}} + \sqrt{\Delta m_{21}^2 + m_{\text{least}}^2} + \sqrt{\Delta m_{32}^2 + m_{\text{least}}^2}, \quad (2.18)$$

<sup>488</sup> where it is clear that a measurement of  $\Sigma_{m_\nu}$  effectively sets the neutrino mass scale  
<sup>489</sup> through  $m_{\text{least}}$ . The analogous formula for the inverted mass ordering is

$$\Sigma_{m_\nu} = m_{\text{least}} + \sqrt{-\Delta m_{32}^2 + m_{\text{least}}^2} + \sqrt{-\Delta m_{31}^2 + m_{\text{least}}^2}. \quad (2.19)$$

<sup>490</sup> In figure 2.3 we plot the observable  $\Sigma_{m_\nu}$  as a function of  $m_{\text{least}}$ .

<sup>491</sup> Upcoming experiments are planned to refine measurements of the CMB, LSS, and  
<sup>492</sup> other cosmological observables. With this additional data it is possible that in the  
<sup>493</sup> near future cosmological measurements will be able to positively constrain the neutrino  
<sup>494</sup> absolute mass scale. However, the strength of these limits strictly depend on the accuracy  
<sup>495</sup> of the  $\Lambda$ CDM model, which highlights the need for direct experimental measurements of  
<sup>496</sup> the neutrino mass to confirm the predictions of cosmology and to fix the neutrino mass  
<sup>497</sup> parameter in future cosmological analyses.

### <sup>498</sup> 2.5.2 Limits from Neutrinoless Double Beta-decay Searches

<sup>499</sup> If neutrinos are Majorana fermions then the neutrino is equivalent to its own antiparticle  
<sup>500</sup> and lepton conservation is not an exact law of nature. Searches for lepton number  
<sup>501</sup> violation, specifically the neutrinoless double beta-decay ( $0\nu\beta\beta$ ) process, are some of the  
<sup>502</sup> most powerful tests of lepton number conservation, which depend on the neutrinos being  
<sup>503</sup> Majorana fermions. In double beta-decay two neutrons contained in the decay species  
<sup>504</sup> nucleus spontaneously decay into two protons resulting in the production of two electrons  
and two neutrinos (see Figure 2.4). However, for  $0\nu\beta\beta$  the two neutrinos self-annihilate

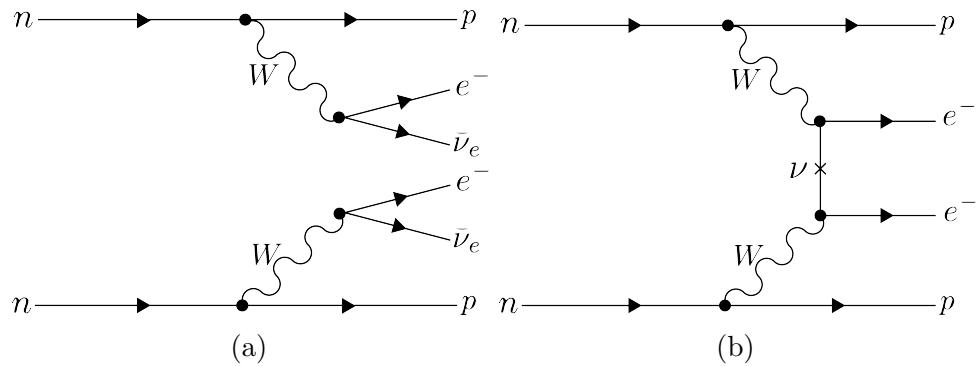


Figure 2.4: Feynman diagrams for double beta-decay (a) and  $0\nu\beta\beta$  (b).

<sup>505</sup>

506 during the decay resulting only in the production of two electrons and a violation of  
 507 lepton number by two.

508 Assuming that the exchange of two Majorana neutrinos is the dominant channel for  
 509  $0\nu\beta\beta$ , then a measurement of the  $0\nu\beta\beta$  half-life for a particular isotope can be used to  
 510 set the neutrino absolute mass scale. The half-life is written in terms of the effective  
 511 neutrino mass for  $0\nu\beta\beta$  ( $m_{\beta\beta}$ ) using the equation

$$T_{1/2}^{0\nu} = \frac{1}{G|\mathcal{M}|^2 m_{\beta\beta}^2}, \quad (2.20)$$

512 where  $G$  is the phase-space factor for the decay and  $\mathcal{M}$  is the relevant nuclear matrix  
 513 element.  $m_{\beta\beta}$  is given by an incoherent sum of the neutrino mass eigenstates weighted  
 514 by the PMNS mixing matrix parameters,

$$m_{\beta\beta} = \left| \sum_i U_{ei}^2 m_i \right|. \quad (2.21)$$

515 The information provided from  $0\nu\beta\beta$  on the neutrino mass scale can be visualized  
 516 by expressing the value of  $m_{\beta\beta}$  in terms of  $m_{\text{least}}$  and two relative Majorana phases.  
 517 The allowed regions for  $m_{\beta\beta}$  as a function of  $m_{\text{least}}$  are shown in Figure 2.5 as the  
 518 regions bounded by the black curves overlayed with the discovery probabilities of future  
 $0\nu\beta\beta$  decay experiments based on current neutrino data. Because of the possibility

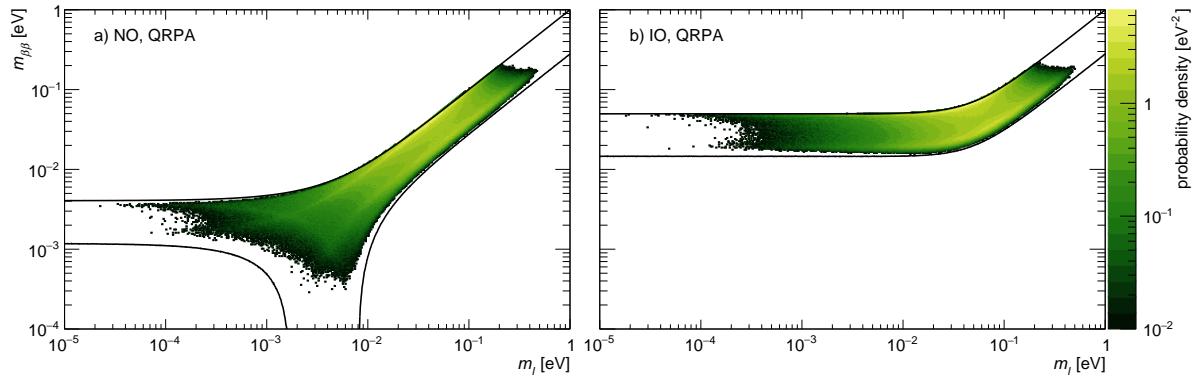


Figure 2.5: The discovery probabilities for the future generation of  $0\nu\beta\beta$  experiments as  
 a function of  $m_{\beta\beta}$  and  $m_{\text{least}}$ .

519  
 520 of cancellation due to the unknown Majorana phases included in the sum specified by  
 521 Equation 2.21, the information gained is necessarily imperfect. Additionally, theoretical  
 522 uncertainty in the calculation of the nuclear matrix elements complicates the calculation  
 523 of  $m_{\beta\beta}$  from a measurement of  $0\nu\beta\beta$  half-life. Similar to cosmology there is a high degree

524 of complementarity between direct measurements of the neutrino mass and  $0\nu\beta\beta$ . In  
 525 particular, a measurement of  $m_{\text{least}}$  to less than than 0.1 eV sensitivity provides significant  
 526 information for  $0\nu\beta\beta$  searches based of the discovery probabilities of Figure 2.5.

### 527 2.5.3 Limits from Beta-decay

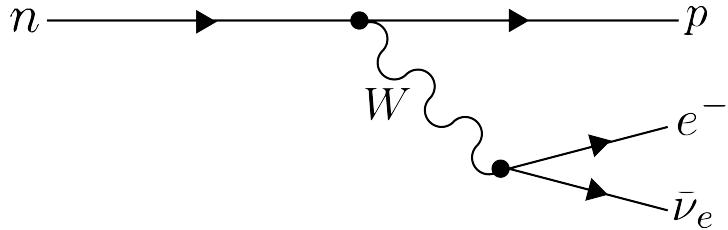


Figure 2.6: A Feynman diagram of beta decay

528 Certain processes involving neutrinos, in particular beta-decay (see Figure 2.6), have  
 529 initial states with well-defined total energies and final states where the kinetic energies  
 530 of each of the particles can be measured with high accuracy and precision. Beta-decay  
 531 involves the decay of an unstable isotope where a neutron spontaneously converts to  
 532 a proton and emits and electron and anti-neutrino ("neutrino" for brevity) to conserve  
 533 charge and lepton number. Therefore, by applying the principles of energy and momentum  
 534 conservation a measurement of the kinematics of the final state can be used to constrain  
 535 the neutrino mass as proposed by Fermi in his 1934 description of nuclear beta-decay  
 (see Figure 2.7). Because the constraint on the neutrino mass from beta-decay depends

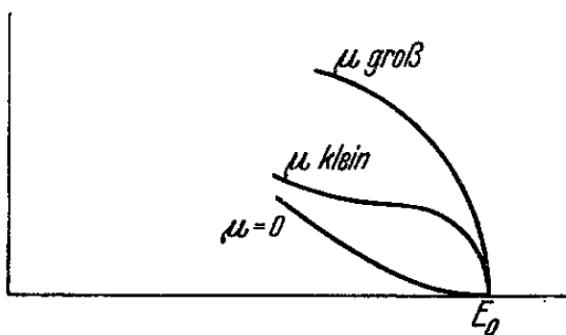


Figure 2.7: A figure from Fermi's 1934 paper on a theory of beta-decay depicting the kinetic energy spectrum of the emitted electron. The effect of the neutrino mass, written as  $\mu$ , is to distort the shape of the spectrum near the endpoint from the zero-mass spectrum.

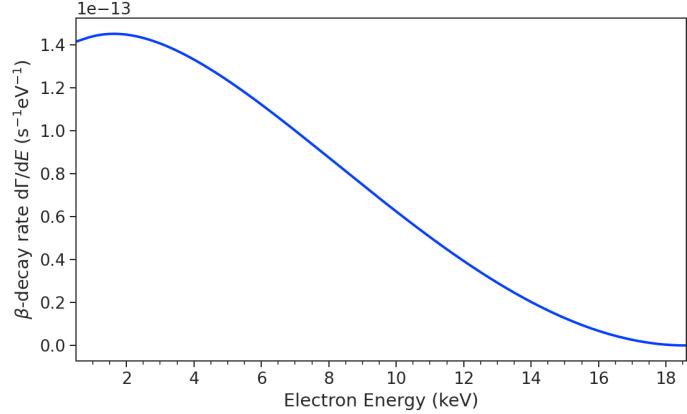
only on the final state measurement capabilities, such measurements of the neutrino mass are often referred to as model-independent or direct in contrast to constraints on the neutrino mass from cosmology and  $0\nu\beta\beta$ .

The beta-decay isotope of choice for direct neutrino mass measurements with beta-decay has been tritium ( $^3H_2$ ) for many decades, because it conveniently fulfills many experimental requirements. Of upmost importance is a decay with a low Q-value, which is the available kinetic energy based on the mass difference between the initial and final states. The effect of a massive neutrino on the shape of the spectrum is magnified for low Q-values and tritium decays have an unusually low Q-value of 18.6 keV. Additionally, tritium beta-decay is a super-allowed decay, which means that it has a relatively short half-life of 12.3 years making it easy to obtain a high-activity source with a relatively small source mass. High-activity is desireable due to the low-activity near the tritium spectrum endpoint. For tritium beta-decays only a factor of  $3 \times 10^{-13}$  of the decays occur in the last 1 eV of the spectrum. Isotopes with Q-values lower than tritium are known, but this is outweighed by exceedingly long half-lives leading to unobtainable source masses.

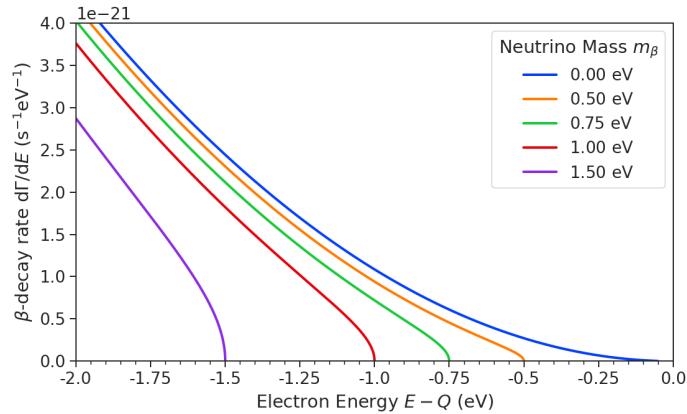
Neutrino mass measurements using beta-decay measure the effect of the neutrino's mass on shape of the electron's kinetic energy spectrum near the endpoint. The kinetic energy spectrum (see Figure ??) is given by

$$\frac{d\Gamma}{dE} = \frac{G_F^2 |V_{ud}|^2}{2\pi^3} (G_V^2 + 3G_A^2) F(Z, \beta) \beta (E + m_e)^2 (E_0 - E) \times \sum_{i=1,2,3} |U_{ei}|^2 [(E_0 - E)^2 - m_i^2]^{1/2} \Theta(E_0 - E - m_i), \quad (2.22)$$

where  $G_F$  is the Fermi coupling constant,  $V_{ud}$  is an element of the CKM matrix,  $E$  is the kinetic energy of the electron,  $\beta$  is the velocity of the electron divided by the speed of light,  $E_0$  is the endpoint energy assuming zero neutrino mass,  $F(Z, \beta)$  is the Fermi function, and  $\Theta(E_0 - E - m_i)$  is the Heaviside function, which enforces energy conservation. One can see that the decay spectrum is actually a combination of three spectra with different endpoints based on the actual values of the neutrino mass eigenstates,  $m_i$ . This results in "kinks" in the spectrum shape due to the overlapping spectra but such an effect would be nearly impossible to resolve given the finite energy resolution of a real experiment.



(a)



(b)

Figure 2.8: The tritium beta-decay spectrum. The affect of a massive neutrino on the spectrum is to change it's shape near the endpoint by an amount proportional to the size of the neutrino mass. This suggests that a sufficiently high-statistic and high-resolution measurement of the spectrum endpoint would be able to measure the neutrino mass.

565

The neutrino mass scale variable measured by beta-decay is given by

$$m_\beta^2 = \sum_i |U_{ei}|^2 m_i^2, \quad (2.23)$$

566 where  $m_{beta}$  is the effective mass of the neutrino in beta-decay or simply neutrino mass  
 567 for brevity. By assuming unitarity, the neutrino mass can be expressed in terms of  
 568 the PMNS matrix elements, squared mass differences, and the lightest neutrino mass

<sup>569</sup> eigenstate. For the normal mass ordering the equation is

$$m_\beta^2 = m_{\text{least}}^2 + |U_{e2}|^2 \Delta m_{21}^2 + |U_{e3}|^2 \Delta m_{31}^2, \quad (2.24)$$

<sup>570</sup> and for the inverted ordering the equation is

$$m_\beta^2 = m_{\text{least}}^2 + |U_{e1}|^2 (-\Delta m_{32}^2 - \Delta m_{21}^2) + |U_{e2}|^2 (-\Delta m_{32}^2). \quad (2.25)$$

<sup>571</sup> Therefore, a measurement of the neutrino mass in combination with neutrino mixing  
<sup>572</sup> parameters is effectively a measurement of  $m_{\text{least}}$ .

<sup>573</sup> Since the neutrino mass is small ( $< 1$  eV), it's effect on the spectrum is limited to  
<sup>574</sup> the endpoint region. The affect of a non-zero neutrino mass on the endpoint spectrum is  
<sup>575</sup> plotted for the reader in Figure 2.8. Resolving the small changes in the spectrum shape  
<sup>576</sup> requires an experimental technique with high statistics, excellent energy resolution, and  
<sup>577</sup> low background activity.

<sup>578</sup> The KATRIN collaboration, utilizing a large MAC-E (magnetic adiabatic collimation  
<sup>579</sup> with electrostactic) filter spectrometer recently obtained the best direct measurement of  
<sup>580</sup> the neutrino mass, with a 90% confidence upper limit of 0.8 eV. With more statistics the  
<sup>581</sup> KATRIN collaboration estimates an ultimate sensitivity to neutrino masses of 0.2 eV.

582 **Chapter 3** |

583 **Direct Measurement of the Neutrino Mass**

584 **with Project 8**

585 **3.1 Introduction**

586 A promising technique for direct measurements of the neutrino mass beyond the projected  
587 limit of the ongoing KATRIN experiment [1] is tritium beta-decay spectroscopy with an  
588 atomic tritium source [2]. Atomic tritium, combined with a large-volume, high-resolution  
589 energy measurement technique, is capable of measuring the neutrino mass with sensitivity  
590 below the 50 meV limit allowed by neutrino oscillations.

591 Cyclotron Radiation Emission Spectroscopy or CRES is a high-resolution energy  
592 measurement technique compatible with atomic tritium production and storage that can  
593 enable the next-generation of neutrino mass direct measurement experiments [3]. The  
594 Project 8 collaboration is currently engaged in a program of research and development  
595 (R&D) aimed at developing the technology necessary for a 40 meV sensitivity measurement  
596 of the neutrino mass using CRES and atomic tritium [4].

597 In Section 3.2 I provide an introduction to the basics of the CRES technique as well as  
598 the goals of the Project 8 experiment. Additionally, I sketch out the phased experiment  
599 development plan being implemented by Project 8 to build towards a next-generation  
600 neutrino mass experiment.

601 In Section 3.3 I give a brief overview of Phase II of the Project 8 experiment [5, 6],  
602 which completed early in 2023. Although the bulk of the work presented in this thesis is  
603 relevant to designs of future Project 8 experiments, a description of the work in Phase II  
604 provides useful context for the rest of the work.

605 In Section 3.4 I introduce a CRES measurement concept based on antenna arrays [7],  
606 which could be the basis for the ultimate Project 8 neutrino mass experiment. A  
607 significant portion of the R&D efforts of Project 8 in Phase III were directed towards

608 simulating and modeling this experimental concept in order to understand the achievable  
609 sensitivity to the neutrino mass.

610 Lastly, in Section 3.5 I introduce conceptual designs of pilot-scale experiments and  
611 Phase IV that combine atomic CRES with a large-volume CRES detection technique.  
612 This includes a design concept for an antenna array based experiment, but also a design  
613 for a resonant cavity based experiment. Resonant cavities are discussed in more depth in  
614 Chapter 6 and have become the default choice for the Phase IV experiment.

## 615 **3.2 Cyclotron Radiation Emission Spectroscopy and Project** 616 **8**

### 617 **3.2.1 Cyclotron Radiation Emission Spectroscopy — CRES**

618 Time and frequency are two of the most precisely measured quantities in physics. It is  
619 often advantageous to convert measurements of other physical quantities like mass or  
620 length into frequency measurements due to the digital nature of frequency measurements  
621 that make them immune to many sources of noise. Atomic clocks, which operate by  
622 measuring the frequencies of various atomic transitions, have been used to measure  
623 time with astounding relative uncertainties of  $10^{-18}$  seconds [8]. The extreme precision  
624 possible with frequency measurements is often summarized using the a quote from the  
625 Physicist Arthur Schawlow who said advise his students to "Never measure anything but  
626 frequency!" [9].

627 Neutrino mass measurements using tritium beta-decay require us to measure pertur-  
628 bations of the 18600 eV tritium endpoint to a precision as low as 0.1 eV, therefore, a  
629 spectroscopic technique with extremely high resolution is required for this measurement.  
630 The intuitive explanation for why frequency measurements are capable of such high reso-  
631 lutions is that they are essentially counting measurements, which average the number of  
632 oscillations of a physical system over time. By observing a rapidly oscillating system over  
633 a sufficient length of time one can obtain essentially arbitrary precision on a frequency  
634 limited only by the time available for measurement and the SNR of the system.

635 What is required is that one translate the kinetic energy of the electron into a frequency,  
636 and a straightforward way to accomplish this is to place a gaseous supply of tritium into  
637 a magnetic field. When an atom decays the resulting electron will immediately begin  
638 to orbit around a magnetic field line at the cyclotron frequency, which is proportional  
639 to its kinetic energy (see Figure 3.1). The acceleration caused by the orbit leads to the

640 emission of cyclotron radiation that can be detected using an array of antennas or a  
 641 different RF sensor such as a resonant cavity. The frequency of the radiation gives the  
 642 electron's kinetic energy, which is used to build the beta-decay spectrum and measure  
 643 the neutrino mass. The name for this measurement technique is Cyclotron Radiation  
 644 Emission Spectroscopy or CRES [3].

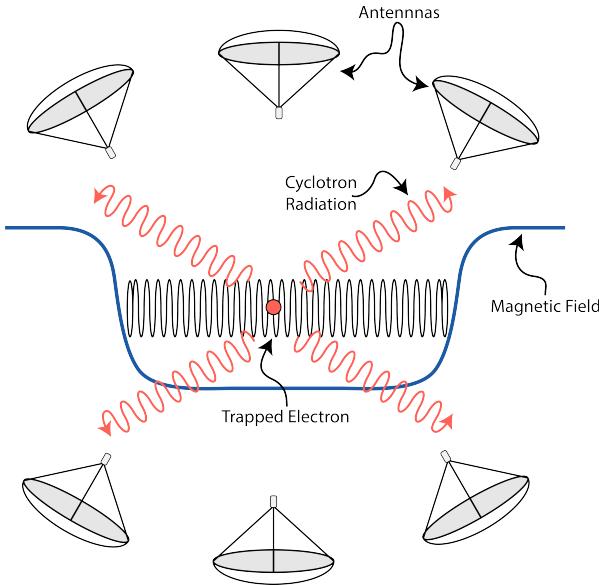


Figure 3.1: A cartoon illustration of the CRES technique. An electron is contained in a magnetic trap so that its cyclotron radiation can be detected by an array of antennas. Detecting the cyclotron radiation allows us to measure its cyclotron frequency and determine its kinetic energy.

645 For non-relativistic particles the cyclotron frequency is simply a function of the  
 646 charge-to-mass ratio of the particle, however, from the relativistic form of the cyclotron  
 647 frequency

$$f_c = \frac{qB}{2\pi m_e \gamma} = \frac{1}{2\pi} \frac{qB}{m_e + E_{\text{kin}}/c^2}, \quad (3.1)$$

648 one can see that the kinetic energy ( $E_{\text{kin}}$ ) of the electron is directly proportional to the  
 649 inverse of the cyclotron frequency ( $f_c$ ). Electrons with kinetic energies of 18.6 keV are in  
 650 the weakly relativistic regime with  $\beta = \frac{v}{c} = 0.263$  and  $\gamma = 1.036$ .

651 The required frequency resolution needed for neutrino mass measurement can be  
 652 obtained by differentiating Equation 3.1,

$$\frac{df_c}{dE_{\text{kin}}} = \frac{1}{2\pi} \frac{-qBc^2}{(m_e c^2 + E_{\text{kin}})^2}, \quad (3.2)$$

653 from which we can obtain the relationship between fractional differences in energy and  
654 frequency,

$$\frac{df_c}{f_c} = \frac{1 - \gamma}{\gamma} \frac{dE_{\text{kin}}}{E_{\text{kin}}}. \quad (3.3)$$

655 Therefore, an energy precision of 1 eV for an 18.6 keV electron requires a frequency  
656 precision of approximately 2 ppm.

657 The minimum observation time required to achieve this resolution can be estimated  
658 using the uncertainty principle as formulated by Gabor [10]. Electrons from tritium  
659 beta-decay experience random collisions with the background gas particles, which limits  
660 the uninterrupted radiation lifetime. The time between collision events, referred to  
661 as track length in the context of CRES measurements, is an exponentially distributed  
662 variable. Differences in the track lengths of a population of mono-energetic electrons leads  
663 to uncertainty or broadening in the distribution of measured frequencies proportional to  
664 the mean track length,  $\tau_\lambda$ . The resulting frequency distribution has a Lorentzian profile,  
665 whose width is given by the Gabor limit,

$$\tau_\lambda \Delta f_c = \frac{1}{2\pi} \implies \Delta f_c = \frac{1}{2\pi\tau_\lambda}. \quad (3.4)$$

666 The cyclotron frequency for a 18.6-keV electron in a 1 T field is approximately  
667 27 GHz, from which one can estimate the minimum observation time for 2 ppm frequency  
668 resolution at approximately 3  $\mu$ sec. The Gabor limit is not the true lower bound on the  
669 frequency resolution for a CRES signal, since it is based on the details of the Fourier  
670 representation of a time-series with a fixed length. If one takes the approach of fitting  
671 the CRES signal in the time-domain, then one finds that the limit on frequency precision  
672 is given by the Cramér-Rao lower bound (CRLB) [11], which depends on the track length  
673 and SNR. The CRLB allows for better precision on the cyclotron frequency, however,  
674 the Gabor limit provides an intuitive limit with the correct order of magnitude.

675 Ensuring that an electron remains under observation long enough so that its frequency  
676 can be precisely measured requires a magnetic trap. A magnetic trap is a local minimum  
677 in a background magnetic field generated an appropriate configuration of electromagnetic  
678 coils. Since magnetic fields can do no work, there is no danger of the magnetic trap  
679 affecting the kinetic energy electron after it is emitted from the beta-decay. One common  
680 approach to creating a magnetic trap is the "bathtub" trap configuration, which in it's  
681 simplest form consists of two high magnetic field pinch coils aligned on a central axis  
682 that are well separated (see Figure 3.2). This configuration produces a trap with a flat  
683 uniform bottom and relatively steep walls, which is ideal for CRES measurements.

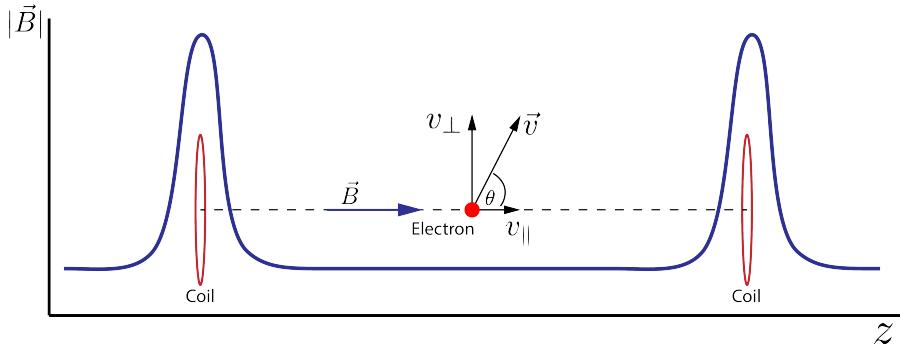


Figure 3.2: An illustration of an electron in a bathtub magnetic trap generated by two well-separated coils.

684 Electrons produced in the trap oscillate back and forth between the trap walls at  
 685 a frequency that depends upon the pitch angle, unless they are produced with pitch  
 686 angles too small to be contained in the trap. Pitch angle is defined as the angle between  
 687 the component of the electron's velocity perpendicular to the magnetic field and the  
 688 component parallel to the magnetic field,

$$\tan \theta = \frac{v_{\perp}}{v_{\parallel}}. \quad (3.5)$$

689 The axial motion of the electron leads to variation in the cyclotron frequency due to  
 690 the changing value of the magnetic fields. This leads to frequency modulation that  
 691 generate sidebands in the cyclotron radiation spectrum. Resolving these sideband  
 692 frequency components is necessary for a complete reconstruction of the CRES signal in  
 693 the experiment.

694 Electrons trapped in a cylindrically symmetric trap have three primary components of  
 695 motion (see Figure 3.3). The dominant component, typically with the highest frequency,  
 696 is the electron's cyclotron orbit, which encodes information on the electron's kinetic  
 697 energy. Axial motion from the electron's pitch angle leads to frequency modulation but  
 698 also a shift in the average magnetic field experienced by an electron. This leads to a  
 699 correlation between the kinetic energy of the electron and the pitch angle depending on  
 700 the particular shape of the magnetic trap, which can negatively impact energy resolution.  
 701 To reduce this correlation one must engineer the trap to have a flat bottom with very  
 702 steep wall both of which are more easily achieved with a small aspect ratio bathtub trap.  
 703 Radial gradients in the trap leads to a third component of motion called grad-B drift [12].

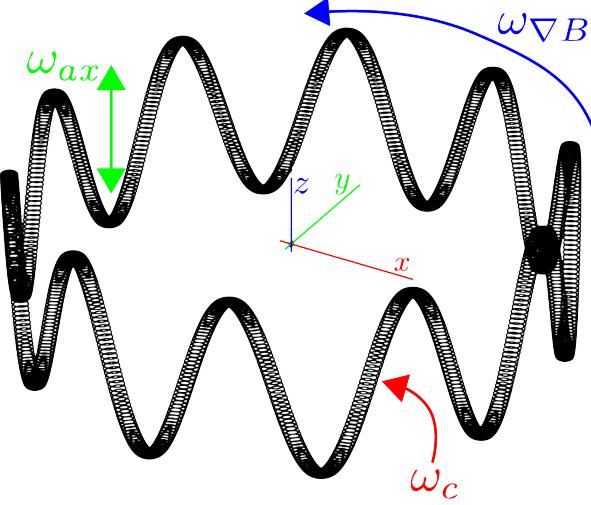


Figure 3.3: A plot of the main components of an electron's trajectory in a cylindrically symmetric trap.

704 The equation for the drift velocity is

$$\mathbf{v}_{\nabla B} = \frac{m_e v_\perp^2}{2qB} \frac{\mathbf{B} \times \nabla B}{B^2}. \quad (3.6)$$

705 These additional components of motion all influence the shape of the CRES signal so  
 706 modeling their effects is critical to proper measurement of the kinetic energy.

707 The total power of the radiation emitted by an electron in a free-space environment  
 708 is given by the Larmor equation [13]

$$P(\gamma, \theta_p) = \frac{1}{4\pi\epsilon_0} \frac{2}{3} \frac{q^2 \omega_c^2}{c} (\gamma^2 - 1) \sin^2 \theta_p, \quad (3.7)$$

709 where  $\omega_c$  is the cyclotron frequency multiplied by  $2\pi$  and  $\theta_p$  is the pitch angle to distinguish  
 710 it from the spherical angle coordinate. A single electron with a 90° pitch angle and  
 711 18.6 keV of kinetic energy in a 1 T magnetic field emits a total radiation power of 1.2 fW,  
 712 which is quite small compared with typical RF systems, furthermore, one is typically  
 713 only able to receive a fraction of this total power with an antenna or other detection  
 714 system. Therefore, RF systems in CRES experiments must be operated at cryogenic  
 715 temperatures to limit the noise power such that adequate SNR can be achieved for signal  
 716 detection and reconstruction. Alternatively, longer tracks enable detection of weaker  
 717 signals due to the increase in the total signal energy available for the detection algorithm.

### **718 3.2.2 The Project 8 Collaboration**

719 The Project 8 collaboration<sup>1</sup> is a group of institutions in the United States and Germany  
720 aiming to measure the neutrino mass by developing a novel spectrometer technology  
721 based on CRES. In the ultimate Project 8 experiment the CRES technique will be used  
722 to measure the beta-decay spectrum using a large source of atomic tritium sufficient to  
723 achieve the required statistics in the last  $O(10)$  eV of the decay spectrum. Project 8 is  
724 targeting a neutrino mass sensitivity below 50 meV [14], which exhausts the range of  
725 possible neutrino masses under the inverted hierarchy and is a factor of four less than  
726 sensitivity projections for the ongoing KATRIN experiment.

727 Project 8's proposed experiment requires the development of two novel technologies:  
728 the production and trapping of a source of atomic tritium on cubic-meter scales and  
729 technology to enable CRES measurements of individual electrons in the same volume.

### **730 Atomic Tritium**

731 Previous measurements of the tritium beta-decay spectrum for neutrino mass measure-  
732 ments have relied on sources of molecular tritium for their measurements [1, 15, 16] due  
733 to the technical challenges associated with the production and storage of atomic tritium.

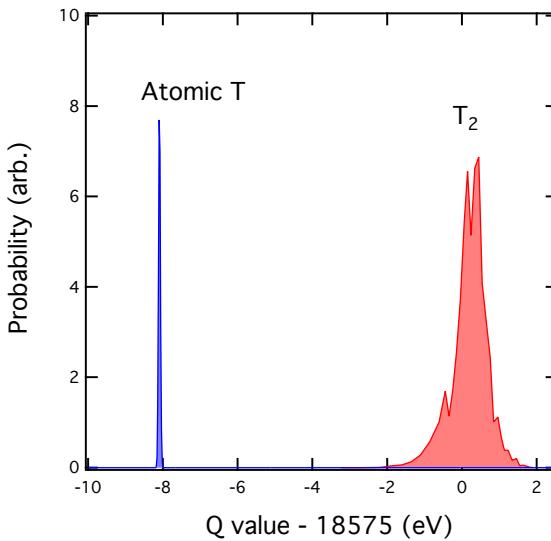


Figure 3.4: A plot of the final state distributions of atomic and molecular tritium. The final state distribution provides the primary contribution to the width of the molecular spectrum whereas thermal doppler broadening is responsible for the width of the atomic spectrum.

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<sup>1</sup><https://www.project8.org/>

734 One must supply sufficient energy to the tritium molecules to break the molecular  
735 bond and create atomic tritium. Common approaches to this include the use of hot  
736 coaxial filament atom crackers as well as plasma atom sources. Both involve heating the  
737 tritium atoms to temperatures of  $> 2500$  K, which must then be cooled to temperatures  
738 on the order of a few mK so that the tritium atoms can be trapped. Cooling the atoms  
739 requires the construction of a large tritium infrastructure and cooling system that can  
740 supply a source of cold atoms to the trap.

741 Once cold tritium atoms are produced they cannot make contact with any surfaces  
742 to avoid recombination of the atoms to molecules. Therefore, a magnetic trap is required  
743 to store the atoms for a sufficient length of time that they have a chance to decay before  
744 escaping the trap. Trapping the atoms requires the construction of a large and complex  
745 magnet system that must be cooled to cryogenic temperatures.

746 The significant experimental complexity caused by atomic tritium makes a molecular  
747 source the obvious choice from practical considerations. However, the drawback of  
748 molecular tritium for neutrino mass measurement is the irreducible broadening in the  
749 electron's kinetic energy due to the final state spectrum of molecular tritium (see Figure  
750 3.4). The broadening of the final state spectra has a RMS amplitude of 436 meV [17, 18]  
751 caused by variation in the final vibrational state of the daughter molecule. For atomic  
752 tritium the primary sources of broadening in the final state spectrum are magnetic  
753 hyperfine splittings (magnitude of  $O(10^{-5})$  eV) and thermal Doppler broadening caused  
754 by the motion of the trapped atom. For atomic tritium at a temperature of 1 mK thermal  
755 broadening is the dominant contribution, providing about 1 meV RMS of broadening to  
756 the electron's kinetic energy.

757 The larger energy broadening with molecular tritium leads to an irreducible statistical  
758 uncertainty that limits the achievable sensitivity to approximately 100 meV at 90%  
759 confidence. For previous direct measurements of the neutrino mass this uncertainty is an  
760 insignificant contribution to the overall uncertainty budget, however, for experiments  
761 like Project 8 atomic tritium is a key component to the success of the experiment.

## 762 **CRES for Neutrino Mass Measurement**

763 Several features of the CRES technique make it an attractive choice for a next generation  
764 neutrino mass measurement experiment. For example, with a CRES experiment the  
765 volume of the source gas can be the same as the volume of the CRES spectrometer.  
766 This is due to the fact that CRES is a remote-sensing technique that can observe the  
767 energy of the electron without altering its trajectory or directly interacting with the

768 electron. Given that tritium gas is transparent to cyclotron radiation the kinetic energies  
769 of electrons can be measured with an appropriate sensing technology, such as a cavity or  
770 antenna array, located directly outside the atom trapping volume.

771 The current state-of-the-art tritium beta-decay spectroscopy experiment, KATRIN,  
772 utilizes the magnetic adiabatic collimation with an electrostatic filter (MAC-E filter)  
773 technique to measure the beta-decay spectrum of molecular tritium. In this approach,  
774 a source of molecular tritium is located outside the spectrometer. When a beta-decay  
775 occurs the electron must exit the tritium source and travel through the MAC-E filter  
776 before it can be detected on the other side of the filter using a charge sensor. The  
777 measurement statistics of the MAC-E filter are limited by the transverse areas of the  
778 tritium source and the filter due to the need to travel through the experiment without  
779 scattering. This scaling is less favorable than the volumetric scaling of CRES due to the  
780 ability to co-locate source and detector.

781 Another promising aspect of the CRES technique is the inherently high precision  
782 of frequency based measurements. The endpoint of the molecular tritium beta-decay  
783 spectrum is approximately 18.6 keV, which dwarfs the neutrino mass scale of  $< 1 \text{ eV}/c^2$   
784 by at least a factor of  $10^5$ . Measuring the effect of such a small mass on a high energy  
785 electron requires excellent energy resolution. Since frequency measurements are essentially  
786 counting measurements they are intrinsically quite accurate due to the ability to measure  
787 the cyclotron frequency by effectively averaging over millions of cyclotron orbits. Using  
788 off-the-shelf RF components its is possible to achieve part-per-million accuracy on the  
789 kinetic energy with the CRES technique.

790 CRES is also nearly immune to typical sources of backgrounds that plague other  
791 experiments. Since CRES operates via non-destructive measurements of the electron's  
792 cyclotron frequency potential sources of background electrons are effectively filtered out  
793 by limiting the frequency bandwidth of the measurement. The fiducial volume of the  
794 experiment is free from any surfaces that could introduce stray electrons and electrons  
795 from sources outside the fiducial volume can be prevented from entering the experiment.

## 796 Neutrino Mass Sensitivity Goals

797 Project 8's ultimate goal is to combine CRES with atomic tritium to measure the neutrino  
798 mass with 40 meV sensitivity at the 90% confidence level (see Figure 3.5). This sensitivity  
799 is sufficient to fully exhaust the range of allowable neutrino masses under the inverted  
800 neutrino mass ordering regime and is approximately an order of magnitude less than the  
801 projected final sensitivity of the KATRIN experiment. Excluding the full neutrino mass

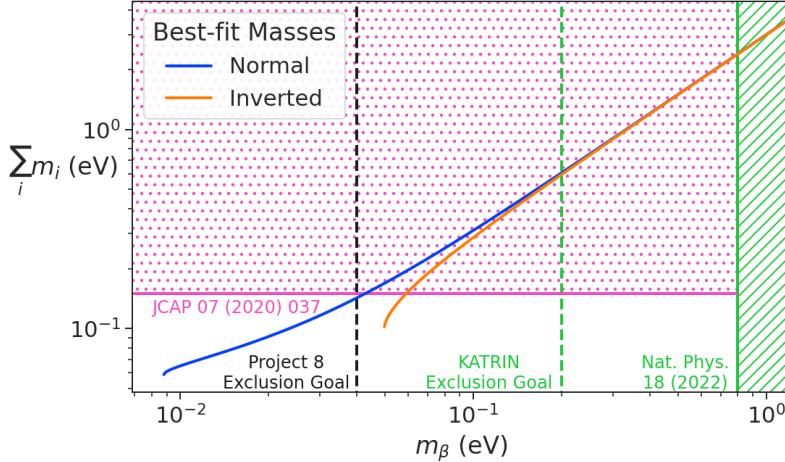


Figure 3.5: Neutrino mass exclusion plot including limits from cosmological measurements and the KATRIN experiment. Allowed ranges for neutrino masses under the normal and inverted hierarchies are shown as the blue and orange lines respectively. The black dashed line shows Project 8’s goal neutrino mass sensitivity for the Phase IV experiment.

parameter space would require a sensitivity an order of magnitude lower than what is proposed by Project 8, which would require an experiment whose size and complexity are currently well beyond proposals for the next-generation of neutrino mass direct measurement experiments.

### 3.2.3 The Project 8 Phased Development Plan

Reaching 40 meV sensitivity requires the simultaneous development and eventual combination of CRES and atomic tritium. These technologies require a significant up-front research and development (R&D) investment to build-out the required capabilities for a 40 meV CRES experiment. Therefore, Project 8 is following a phased experiment plan in which incremental progress can be made towards the ultimate goal of a 40 meV neutrino mass measurement with CRES.

#### Phase I and II: Proof of Principle and First Tritium Measurements

The earlier phases of the Project 8 experiment, Phase I and II, were focused on demonstration and development of the CRES technique itself as well as a proof-of-principle measurement of the neutrino mass using the CRES technique.

In Phase I, Project 8 performed a proof-of-principle measurement of the  $^{83m}\text{Kr}$

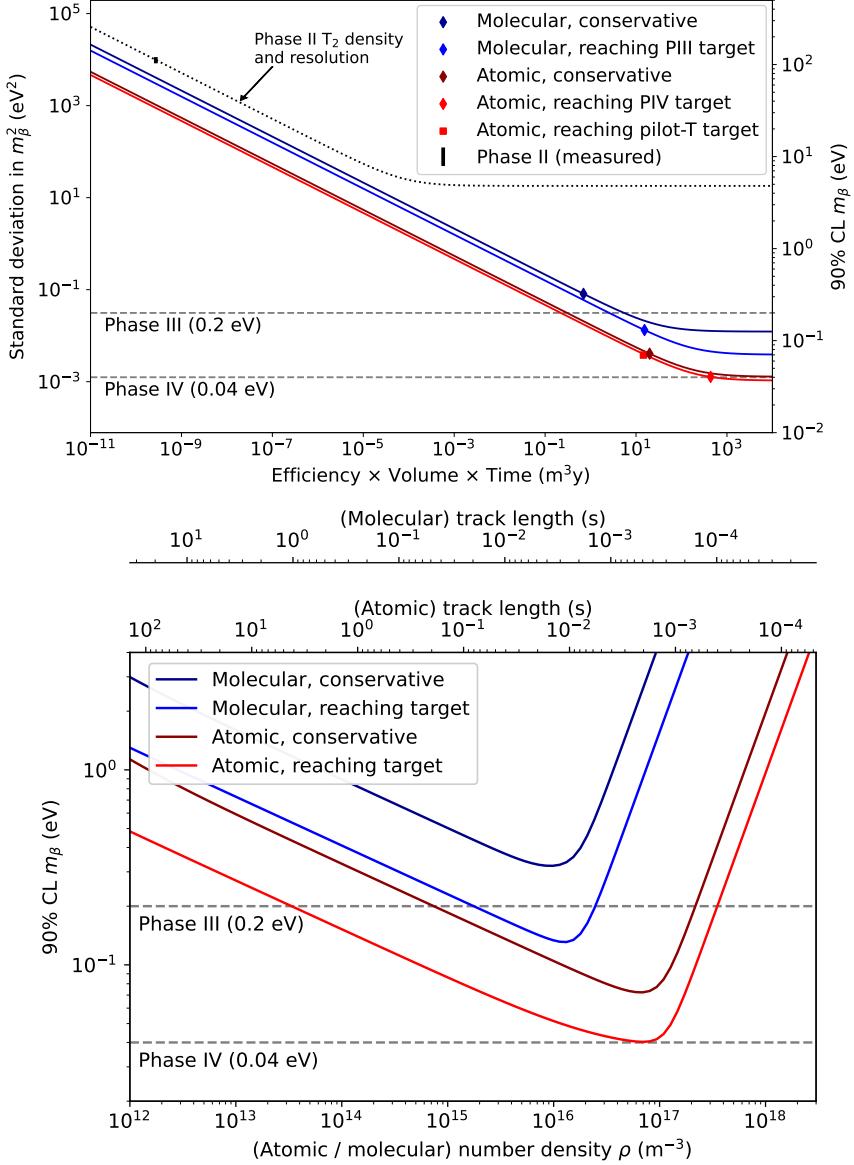


Figure 3.6: Sensitivity calculations for a cavity based CRES experiment that demonstrate the neutrino mass measurement goals of the Project 8 collaboration throughout the phased development plan. The blue curves indicate molecular tritium sources and the red curves indicate atomic tritium sources. In the current plan Phase III contains two tritium experiments. The first is the Low-frequency Apparatus (LFA) which is a molecular tritium experiment and the second is the atomic tritium pilot-scale experiment that ends Phase III. The sensitivity of these experiments is primarily a function of statistics, however, there is a critical density beyond which CRES electrons do not have enough time to radiate between collisions for a high-resolution frequency measurement leading to worse sensitivity.

spectrum using CRES, which marked the first ever energy spectrum measurement with CRES. The experiment included all of the main components expected for the full-scale version of the experiment. An electron source consisting of a gas of  $^{83m}\text{Kr}$  was supplied to a waveguide gas cell constructed out of a segment of WR-42 waveguide and sealed with Kapton windows at the top and bottom. A magnetic trapping region was created in the waveguide cell using a single electromagnetic coil wrapped around the waveguide which provided a trapping volume on the order of a few cubic-millimeters. Detection of the cyclotron radiation was performed by connecting the waveguide cell to an additional segment of waveguide that transmitted the radiation to a cryogenic amplifier.

Success in Phase I was achieved with the 2014 publication of the measured  $^{83m}\text{Kr}$  conversion spectrum [19], which contains a mono-energetic 17.8-keV as well as several other conversion lines at higher energies. Publication of this result marked the official end of Phase I and the start of Phase II in which Project 8 shifted its focus to the demonstration of the first tritium beta-decay spectrum using CRES. Phase II successfully concluded in 2023 with the submission of the papers demonstrating the first tritium beta-decay spectrum endpoint and neutrino mass measurement using CRES. For more information on Phase II please see Section 3.3.

### Phase III: Research and Development and a Pilot-scale Experiment

After Phase II Project 8 has shifted focus to R&D towards the construction of an experiment that demonstrates all the technologies required for a 40 meV measurement of the neutrino mass. The goal for this pilot-scale experiment is to successfully retire all technological and engineering risks associated with the Phase IV experiment, while being a scientifically interesting experiment in its own right that has sensitivity to neutrino masses on par with KATRIN’s final projected sensitivity.

Phase III R&D is divided into two equally important efforts — atomic tritium and CRES detection techniques. Atomic tritium development in Phase III includes the development of all aspects of the tritium system. This includes the production of tritium atoms, atomic cooling and recirculation systems, purity and isotope concentration monitoring, and atom trapping. Currently, Project 8 is operating small scale demonstrator systems developing atom crackers to show that atom production at the estimated rates needed for Phase IV is achievable. Future efforts will continue the current developments on atom production and expand to include demonstrations of atomic cooling with an evaporative beam line as well as atom trapping using Halbach magnet arrays.

The need for new CRES detection techniques is driven by the drastic increase in scale

from Phase II to the Phase IV and the pilot-scale experiments. The physical volume used for CRES in Phase II was on the order of a few cubic-centimeters, and achieving Project 8’s sensitivity target of 40 meV requires an experiment volume on the multi-cubic meter scale. Therefore, the waveguide gas cell CRES detection technique used in Phase II is not a feasible option for the future of Project 8 due to it’s inability to scale to the required size.

Two alternative CRES detection techniques have been proposed for the pilot-scale experiment — antenna arrays and resonant cavities (see Section 3.4 and Chapter 6). Both approaches have relative advantages and disadvantages, however, the improved understanding of the antenna array and cavity approaches to CRES in the recent years has led to cavities being the preferred technology for the pilot-scale experiment and Phase IV due to the estimated reduced cost and complexity of this approach. Since a large degree of the work presented in this thesis is focused on the development of the antenna array CRES technique as well as the design of demonstrator experiments, we described the proposed R&D plan for antenna array CRES in Phase III in Section 3.4.

Cavity CRES R&D in Phase III consists of a series of demonstrator experiments intended to demonstrate cavity CRES at a variety of scales and magnetic fields using electrons from  $^{83m}\text{Kr}$ , an electron gun, and potentially molecular tritium sources. The near-term cavity effort in Project 8 is the cavity CRES apparatus (CCA), which is a small-scale cavity experiment operating near 26 GHz, that will perform the first CRES measurements using a small cavity. This experiment will pave the way towards larger scale cavity experiments in preparation for the eventual pilot-scale tritium experiment.

The pilot-scale experiment is the first experiment, which will combine atomic tritium and large-volume CRES detection in the same experiment. It will directly demonstrate all the technologies required for Phase IV such that no technical risks remain for scaling the experiment to required scale. A robust approach to scaling the pilot-scale experiment is to simply build multiple copies of it for the Phase IV experiment.

## Phase IV: Project 8’s Ultimate Neutrino Mass Experiment

The design of Phase IV should be a direct extension of the pilot-scale CRES experiment that marks the official end of Phase III (see Section 3.5). The Phase IV experiment represents the final experiment in the Project 8 neutrino mass measurement experiment plan and will have sensitivity to neutrino masses of 40 meV.

884 **3.3 Phase II: First Tritium Beta Decay Spectrum and**  
885 **Neutrino Mass Measurement with CRES**

886 In Phase II Project 8 demonstrated the first ever measurement of the tritium beta-decay  
887 spectrum endpoint using the CRES technique, which lead to the first neutrino mass  
888 measurement by the Project 8 collaboration. This milestone was made possible by many  
889 improvements in the CRES technique and in the understanding of CRES systematics,  
890 which takes an important first step towards larger scale measurements of the tritium  
891 beta-decay spectrum with CRES. In this section, I briefly describe some important  
892 elements of the Phase II experiment, with the goal of contextualizing the research and  
893 development efforts for Phases III and IV of Project 8. For more complete descriptions of  
894 the work that lead to Project 8’s Phase II results please refer to the relevant publications  
895 by the collaboration [5, 6].

896 **3.3.1 The Phase II CRES Apparatus**

897 **Magnet and Cryogenics**

898 The magnetic field for the the Phase II experiment is provided by a nuclear magnetic  
899 resonance (NMR) spectroscopy magnet with a central bore diameter of 52 mm (see Figure  
900 3.7). The magnet produces a background magnetic field with an average value of 0.959 T  
901 and a 10 ppm variation across the bore diameter achieved using several shim coils built  
902 into the magnet. Using an external NMR field probe the variation of the magnetic field  
903 along the vertical axis of the magnet bore was measured to obtain an accurate model of  
904 the magnetic field so that the CRES cell could be positioned for optimal magnetic field  
905 uniformity.

906 An external solenoid magnet was installed inside the magnet bore to provide the  
907 ability to shift the magnitude of the background magnetic field by values on the order of  
908 a few mT. The solenoid has inside diameter of 46 mm and a length of 350 mm, which  
909 terminates in a vacuum flange that allows it to be inserted into the NMR magnet bore  
910 from the bottom. By shifting the value of the magnetic field by a few mT, the cyclotron  
911 frequencies of electrons produced by the 17.8 keV  $^{83m}\text{Kr}$  internal-conversion line [20]  
912 can be shifted over a range of frequencies on the order of 100 MHz. This allows one to  
913 study the frequency dependent behavior of multiple CRES systematics such as detection  
914 efficiency that directly affect the measured shape of the tritium spectrum.

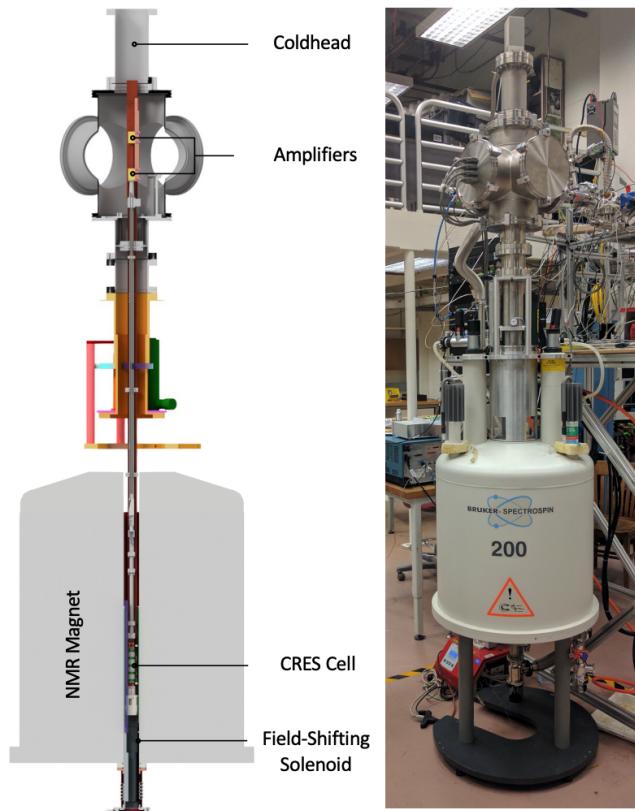


Figure 3.7: The Phase II CRES apparatus used to perform the first measurement of the tritium beta-decay spectrum using CRES.

915        The inside of the magnet bore diameter was pumped down to a vacuum of less than  
 916      10  $\mu\text{torr}$  using a turbomolecular pump, which allows for cryogenic cooling of the CRES  
 917      cell and RF system. Cooling power was supplied to the Phase II apparatus using a  
 918      cryopump with its coldhead mounted above the primary magnet and CRES cell. This  
 919      arrangement allowed for sufficient cooling power to be delivered to the amplifiers to cool  
 920      them to a temperature of  $\approx 40$  K, while keeping the amplifiers far enough from the  
 921      magnet so as not to be damaged by the large field strength. Thermal contact between  
 922      the coldhead, amplifiers, RF system, and CRES cell is achieved using a copper bar that  
 923      runs the full length of the apparatus. To prevent freeze-out of  $^{83m}\text{Kr}$  on the walls of the  
 924      CRES cell a separate heater was installed to keep the CRES cell near a temperature of  
 925      85 K during the operation of the experiment.

926    **CRES Cell**

927    Located in the most uniform region of the magnetic field is the CRES cell, which is the  
928    region of the apparatus where radioactive decays of  $^{83m}Kr$  and  $T_2$  emit electrons that can  
be trapped and measured using CRES (see Figure 3.8). The CRES cell is manufactured

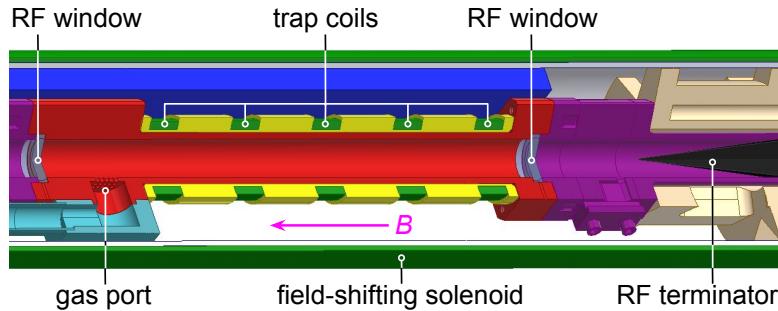


Figure 3.8: Diagram of the CRES cell portion of the Phase II apparatus.

929  
930    from a segment of cylindrical waveguide designed to operate at K-band frequencies  
931    near 26 GHz. The diameter of the waveguide determines which resonant modes of the  
932    waveguide will couple to the electron and transmit its radiation to the amplifiers. For  
933    Phase II a waveguide diameter of 1 cm was selected, which allows electrons to couple to  
934    the TE<sub>11</sub> and TM<sub>01</sub> cylindrical waveguide modes. To reduce complexity in modeling and  
935    analyzing the CRES data, it is ideal to select a diameter that prevents electrons from  
936    coupling to higher-order waveguide modes beyond the fundamental TE and TM modes.

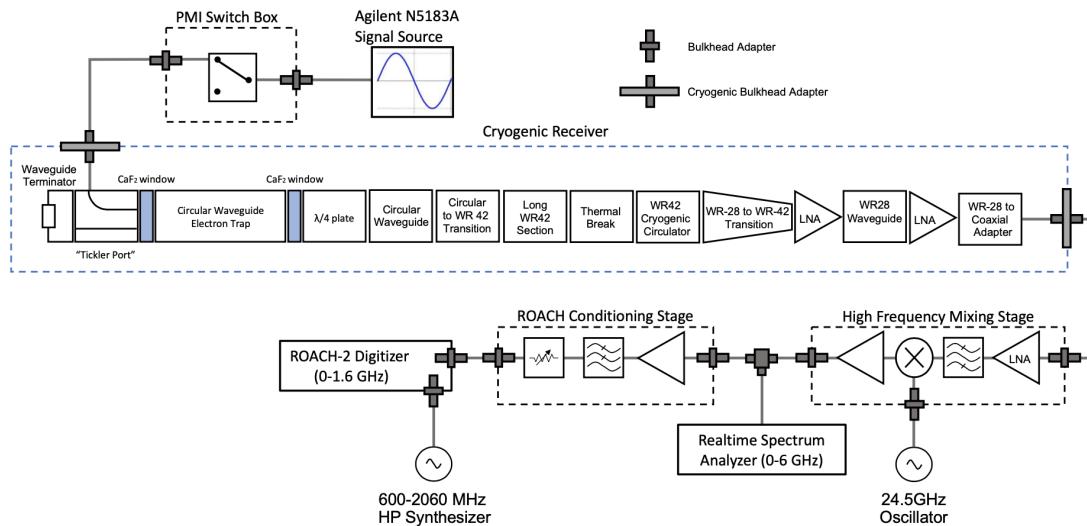
937    Around the exterior of the cylindrical waveguide are several magnetic coils used to  
938    produce magnetic traps inside the CRES cell volume. Without a magnetic trap electrons  
939    produced from decays inside the CRES cell quickly impact the cell wall, which prevents  
940    a measurement of their cyclotron frequency using CRES. Each coil along the length of  
941    the waveguide produces a separate trap that is approximately harmonic in shape. By  
942    independently controlling the currents provided to each coil the traps can be configured  
943    to have equal values of the magnetic field at the trap bottom despite a non-uniform field  
944    from the NMR magnet.

945    Two primary magnetic trap configurations were used during the Phase II experiment.  
946    The first was a shallow trap configuration used primarily for it's high energy resolution to  
947    study systematics using  $^{83m}Kr$  decays, and the second was a deeper trap that could trap a  
948    higher percentage of pitch angles. The trade-off with this trap is that the higher trapping  
949    efficiency comes at the cost of lower energy resolution due to the greater variation in pitch  
950    angle. The deep trap was the trap used to measure the tritium beta-decay spectrum in  
951    Phase II.

952        The source gases were delivered into the CRES cell through a gas port located near the  
 953 top end of the cylindrical waveguide. To prevent the gases from escaping the cell, vacuum  
 954 tight RF transparent windows are needed to contain the tritium and krypton source  
 955 gas across a 1 atm pressure differential, while still transmitting the cyclotron radiation  
 956 without distortion. The crystalline material,  $\text{CaF}_2$ , which has a thermal expansion  
 957 coefficient similar to that of copper, was used for this purpose in the CRES cell. Two  
 958 windows, each 2.4 mm thick, were used to seal off the ends of the CRES cell. The  
 959 thickness of 2.4 mm corresponds to half of a cyclotron wavelength when one accounts for  
 960 the permittivity of  $\text{CaF}_2$ .

## 961        RF System

962        The RF system in the Phase II apparatus transferred the cyclotron radiation from the  
 963 CRES cell to the receiver chain. The receiver chain performs the down-conversion and  
 964 digitization required to obtain signals that can be analyzed to determine the cyclotron  
 frequencies of electrons in the CRES cell (see Figure 3.9).



965        Figure 3.9: RF system diagram for the Phase II apparatus.

966        Below the CRES cell, at the bottom of the Phase II apparatus, is a tickler port and  
 967 waveguide terminator. The tickler port is used to inject signals into the CRES cell and  
 968 RF system for testing and calibration purposes. The waveguide terminator is designed to  
 969 absorb cyclotron radiation emitted by electrons that transmits out of the bottom of the  
 970 CRES cell. This lowers the total power received from electrons in the CRES cell, since all  
 971 the energy radiated downwards is absorbed into the terminator. Earlier iterations of the

972 Phase II apparatus used an RF short in this location that reflected this power up towards  
973 the amplifiers, however, interference between the upward traveling and reflected radiation  
974 led to a disappearance in the signal carrier that made reconstruction impossible.

975 Radiation traveling upward passes through the  $\text{CaF}_2$  window passes through a  $\lambda/4$  plate,  
976 which transforms the circularly polarized cyclotron radiation into linear polarization.  
977 The linearly polarized fields next travel through a segment of circular waveguide that  
978 transitions into a long segment of WR-42 waveguide that carries the fields out of the  
979 high magnetic field region. A thermal break segment is included, which consists of a  
980 segment of gold-plated stainless steel WR-42 waveguide, to help thermally isolate the  
981 relatively warm CRES cell from the colder amplifiers. The radiation then passes through  
982 a cryogenic circular, which prevents signals reflected from the amplifiers from interfering  
983 with the CRES cell before a WR-42 to WR-28 transition connects the waveguide to the  
984 first of the cryogenic amplifiers. The radiation passes through two cryogenic amplifiers  
985 before being coupled to a coaxial termination at the top of the Phase II apparatus.

986 The coaxial cable transfers the cyclotron radiation signals to a high-frequency mixing  
987 stage that performs an analog frequency down-conversion using a 24.5 GHz LO. Two forms  
988 of digitization can be used at this stage to readout the CRES data. One is a real-time  
989 spectrum analyzer that digitizes the CRES signal data in time-domain and computes the  
990 frequency spectrum in real-time, which allows for direct visualization of CRES signal  
991 spectrograms as the experiment is running. The real-time spectrum analyzer is most  
992 useful for taking small amount of streamed data for debugging and analysis of the system.  
993 The other method, which was used to collect the majority of the CRES data in Phase II,  
994 is a ROACH-2 FPGA and digitizer system. The ROACH system consists of a fast ADC  
995 that samples the CRES signal data at 3.2 GSps. Internal digital down-conversion stages  
996 implemented in the FPGA perform a mixing operation that reduces the bandwidth of the  
997 CRES signals to 100 MHz. The FPGA implements a 8192 sample FFT and packetizes  
998 time and frequency domain records in parallel. The packetized data is then transferred  
999 from the ROACH to be analyzed by the data-processing pipeline.

### 1000 **3.3.2 CRES Track and Event Reconstruction**

#### 1001 **Time-Frequency Spectrogram**

1002 The online data-processing is intended to identify interesting data that could contain  
1003 CRES signals using a software real-time triggering algorithm. Interesting segments of  
1004 data identified by this algorithm are collected into files that are transferred to a server for

1005 offline processing and analysis. The data files contain a continuous series of time-domain  
1006 samples, broken into a set of records, which are 4096 samples long. The time-series is  
1007 made up of 8-bit IQ samples acquired at 100 MHz.

1008 Each time-series record is accompanied by an associated frequency spectrum consisting  
1009 of 4096 frequency bins approximately 24.4 kHz wide, which is represented as a power  
1010 spectral density. The individual frequency spectra can be organized temporally to create  
1011 a time-frequency spectrogram that represents the evolution of the cyclotron frequency  
spectrum over the course of the CRES event (see Figure 3.10). The time-frequency

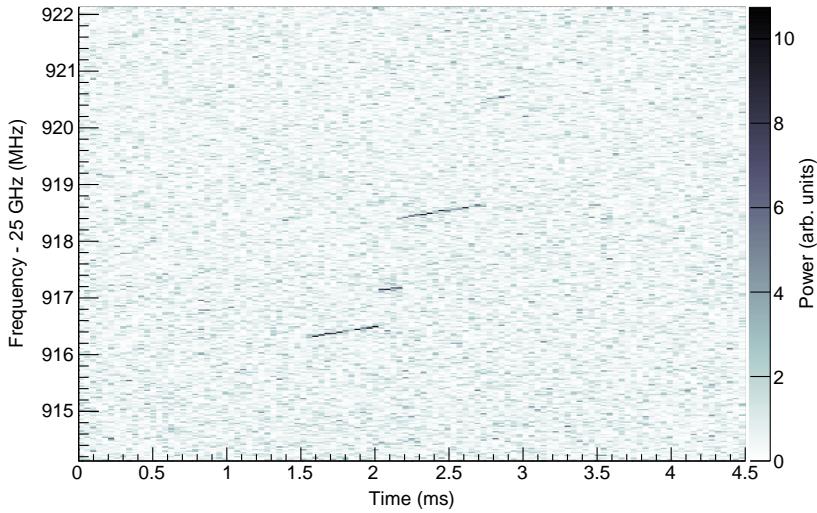


Figure 3.10: The time-frequency spectrogram of a tritium CRES event in the Phase II apparatus.

1012 spectrogram is represented as a two-dimensional image where the color of each pixel is  
1013 proportional to the power spectral density. Each vertical slice of pixels in the image  
1014 represents a frequency spectrum, therefore, each horizontal bin represents the data  
1015 obtained over a duration of  $4096 \times 0.01 \text{ MHz}^{-1} = 40.96 \mu\text{sec}$ .

### 1017 CRES Event Data Features

1018 Phenomenologically, a CRES signal appears as a sinusoidal signal whose frequency slow  
1019 increases ("chirps") over time. Axial motion of the electron in the trap leads to the  
1020 formation of frequency sidebands that surround the more powerful carrier frequency, due  
1021 to Doppler modulation of the electron's frequency as it bounces between the walls of the  
1022 magnetic trap. The critical piece of information that must be extracted from the track  
1023 and event reconstruction procedure is the carrier frequency, since it is this frequency

1024 that gives the cyclotron frequency and thus the kinetic energy. While axial motion from  
1025 non- $90^\circ$  pitch angles does change the average magnetic field experienced by an electron  
1026 and, therefore, changes the cyclotron frequency. Because of low-SNR sidebands were  
1027 unable to be observed in Phase II, so a correction for the effect of the pitch angle on the  
1028 cyclotron frequency was not possible.

1029 In the time-frequency spectrogram representation the chirping carrier frequency  
1030 appears as a linear track of high-power frequency bins (see Figure 3.10). The vertical  
1031 slope of the tracks is caused by the emission of energy from the electron in the form of  
1032 cyclotron radiation, therefore, the size of the slope parameter is directly proportional  
1033 to the Larmour power. The continuous track is periodically interrupted by random  
1034 jumps to higher frequency and lower energy caused by random inelastic collisions with  
1035 background gas molecules. The length of a track is an exponentially distributed variable  
1036 whose mean value is inversely proportional to the gas density. The size of the frequency  
1037 discontinuities is directly proportional to the energies of the rotational and vibrational  
1038 states of background gas molecules.

1039 A CRES event refers to the collection of tracks produced by a trapped electron until  
1040 it inevitably scatters into a pitch angle that can no longer be trapped. The goal of track  
1041 and event reconstruction is to first identify the set of tracks present in a time-frequency  
1042 spectrogram that represents a segment of data acquired in the Phase II apparatus. These  
1043 tracks must then be clustered into events from which we can determine the first track  
1044 produced by the electron and thus estimate it's starting cyclotron frequency and kinetic  
1045 energy.

## 1046 Track Reconstruction

1047 The first step in CRES event reconstruction is the identification of tracks in the time-  
1048 frequency spectrogram, which is essentially an image processing task. Track finding  
1049 starts by normalizing the power spectral density based on the average noise power.  
1050 Next a power threshold is applied to the normalized spectrogram where only bins that  
1051 have a signal-to-noise ratio greater than five are selected to build tracks. In this case  
1052 signal-to-noise ratio is defined as the ratio between the normalized, unitless power of a  
1053 bin divided by the average normalized power across the full frequency spectrum.

1054 The sparse spectrogram produced by this power cut consists only of a sparse collection  
1055 of high-power frequency bins that could be part of a CRES signal track (see Figure  
1056 3.11). In this form is it much easier to identify tracks "by eye", however, for the Phase II  
1057 analysis Project 8 developed its own custom-made track finding algorithm, called the

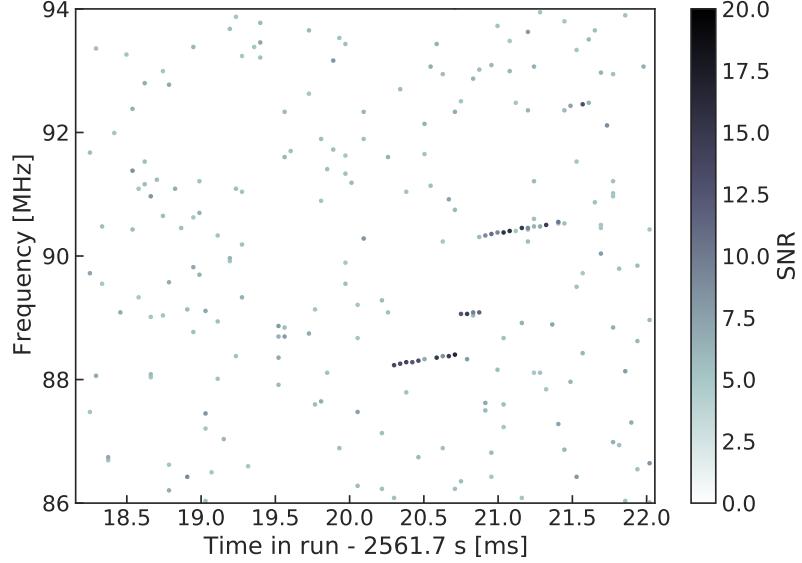


Figure 3.11: The sparse spectrogram obtained by placing a power cut on the raw spectrogram shown in Figure 3.10.

1058 sequential track finder (STF).

1059 The STF algorithm processes the sparse spectrogram in sequential fashion, processing  
 1060 each time-slice one-by-one until the end of the spectrogram is reached. Tracks are found  
 1061 by searching for points in the sparse spectrogram that appear to fall on a straight line.  
 1062 Multiple configurable parameters are built into the STF algorithm that allow the user to  
 1063 tune the criteria for adding a point to an existing track or creating a new track. These  
 1064 include parameters such as maximum time and frequency differences between subsequent  
 1065 points in a track as well as minimum SNR values for the start and endpoints of the track.  
 1066 Additionally, tracks are required to have a minimum length and slope to be considered  
 1067 potential CRES tracks rather than random noise fluctuations.

1068 The resulting output of the STF is a collection of track objects that consist of the track  
 1069 point objects and their properties. The final step is to calculate track-level properties  
 1070 and apply cuts to reject false tracks found by the STF. This involves the fitting of a  
 1071 line to the collection of track points as well as the total and average power of the track  
 1072 obtained by computing the sum and mean of the points powers. The starting frequency  
 1073 of the track is determined by calculating the time coordinate that intersects with the  
 1074 linear fit. A cut is performed to remove all tracks that do not have a specified average  
 1075 power over their duration, which helps to remove the majority of noise fluctuations that  
 1076 have passed all previous cuts up to this point.

1077 **Event Reconstruction**

1078 After track reconstruction comes event reconstruction where the identified tracks are  
1079 grouped into events that correspond to the trajectory of a single electron in the trap. This  
1080 procedure attempts to match tracks head to tail by checking if the start and end times  
1081 of a pair of tracks falls within a certain tolerance. This tolerance is a configurable  
1082 parameter that can be tuned to an optimal value using Monte Carlo simulations of events  
1083 in the Phase II apparatus.

1084 After the event building procedure has completed there remains a small likelihood  
1085 that false tracks have made it through to the event reconstruction stage. Typically, cuts  
1086 at the track level are able to remove 95% of the false tracks identified by the STF, which  
1087 leads to a significant number of false tracks at the event building stage. However, the  
1088 additional event-level information makes it possible to reject events that contain these  
1089 false tracks with a high degree of confidence.

1090 Two event level features are associated with events caused by real electrons — the  
1091 duration of the first track as well as the number of tracks in the event. Real electrons  
1092 tend to have event structures with longer first tracks and a higher number of total tracks.  
1093 Based on the values of these two criteria, a minimum threshold on the average power in  
1094 the first track was configured to reject false events. The average power in the first track  
1095 was chosen due to the critical nature of the starting frequency of the first track in an  
1096 event to the krypton and tritium spectrum analyses.

1097 **3.3.3 Results from Phase II**

1098 The main result from Phase II was the measurement of the tritium beta-decay spectrum  
1099 using CRES, which lead to the first neutrino mass limit with CRES. However, Phase  
1100 II also included a significant  $^{83m}\text{Kr}$  measurement campaign to understand important  
1101 systematics relevant to the tritium spectrum measurement, but also to understanding the  
1102 fundamentals of the CRES technique itself. This required high-resolution measurements  
1103 of the  $^{83m}\text{Kr}$  internal-conversion spectrum [20], which is an interesting science result in  
1104 its own right.

1105 The results from Phase II represents a significant effort from the entire Project 8  
1106 collaboration over several years. Because the focus of my contributions to Project 8 is  
1107 directed towards the research and development efforts for the Phase III experiments, the  
1108 goal in this section is not to provide a detailed description of the analyses that lead to  
1109 the Phase II results. Rather, I will provide brief descriptions of a few plots representative

1110 of the main results from Phase II.

## 1111 Measurements with Krypton

1112 Measurements with krypton were a key calibration tool for Phase II of the experiment and  
1113 will continue to be useful in Phase III. In the context of Project 8 krypton measurements  
1114 refers to CRES measurements of the internal-conversion spectrum of the metastable state  
1115 of krypton-83,  $^{83m}\text{Kr}$ , produced by electron capture decays of  $^{83}\text{Rb}$ . A supply of  $^{83}\text{Rb}$   
1116 was built into the Phase II apparatus gas system that supplied the CRES cell with  $^{83m}\text{Kr}$   
1117 via emanation.

1118 The  $^{83m}\text{Kr}$  internal-conversion spectrum consists of several lines based on the orbital  
1119 of the electron ejected during the decay. The conversion lines useful to Project 8 are  
1120 those that emit electrons with kinetic energies that fall inside the detectable frequency  
1121 bandwidth of the Phase II apparatus. These are the K; L2 and L3; M2 and M3; and N2  
1122 and N3 lines with kinetic energies of 17.8 keV,  $\approx$  30.4 keV,  $\approx$  31.9 keV, and  $\approx$  32.1 keV,  
1123 respectively. The different energies of the lines allow a onw to test the linearity of the  
1124 relationship between kinetic energy and frequency across the range of frequencies covered  
1125 by the continuous tritium spectrum.

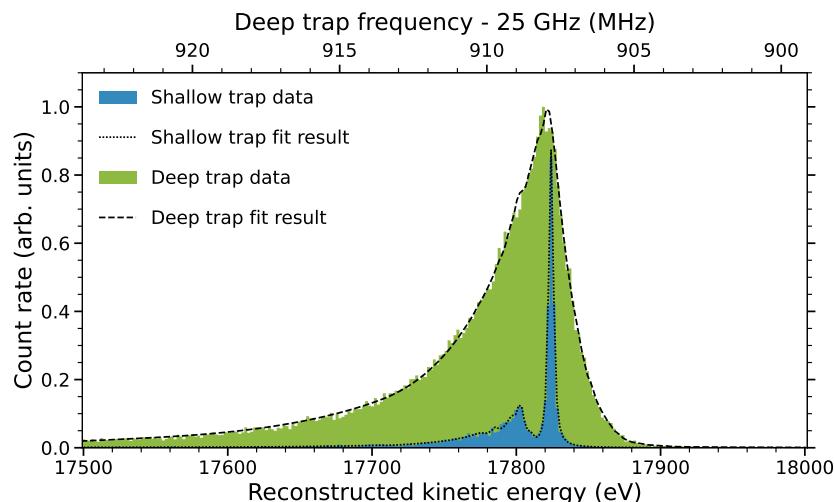


Figure 3.12: Fits to the measured 17.8-keV  $^{83m}\text{Kr}$  conversion line using the deep and shallow trap configurations.

1126 Numerous detector related effects relevant to the tritium analysis can be characterized  
1127 by measuring the shape of the krypton spectrum. Specific examples include variations  
1128 in the magnetic field as a function of the radial position of the electron, variation in  
1129 the magnetic field caused by the trap shape, variation in the average magnetic field for

1130 electrons with different pitch angles, and the effect of missing tracks due to scattering.  
1131 These spectrum shape measurements focused on the 17.8-keV krypton line and utilized  
1132 different trap geometries based on the particular goal of the dataset (see Figure 3.12).

1133 Krypton measurements with a shallow trap allow for high energy resolution, since  
1134 variation in frequency due to pitch angle differences is sharply reduced in the shallow  
1135 trap configuration. With this trap the main 17.8-keV peak of the conversion spectrum is  
1136 clearly visible along with additional satellite peaks at lower energy, which correspond to  
1137 the shakeup/shakeoff spectrum of the decay. The high accuracy of the fit demonstrates a  
1138 high degree of understanding of the CRES systematics.

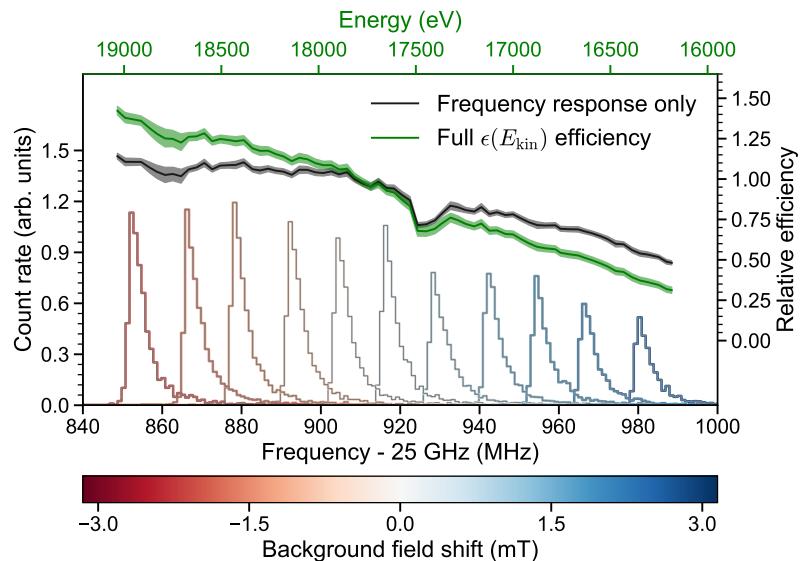


Figure 3.13: Measurements of the 17.8-keV  $^{83m}\text{Kr}$  line using the deep trap configuration for different values of the magnetic field from the field shifting solenoid.

1139 The broadening of the krypton spectrum seen for the deeper track is due to the large  
1140 range of electron pitch angles that can be trapped. Furthermore, with a deeper trap  
1141 there is a larger parameter space of electron that could be produced with pitch angles  
1142 that are trappable but not visible in the time-frequency spectrogram. These electrons  
1143 live in the trap and can scatter multiple times before randomly scattering to a visible  
1144 pitch angle. This leads to one or more missing tracks earlier in the event, which leads to  
1145 a misreconstruction of the true starting frequency. By measuring the krypton spectrum  
1146 shape in the same trap used to detect tritium events, the effect this has on the spectrum  
1147 shape can be characterized to mitigate its impact on the tritium measurements.

1148 Changes in the Krypton spectrum shape as a function of CRES frequency were

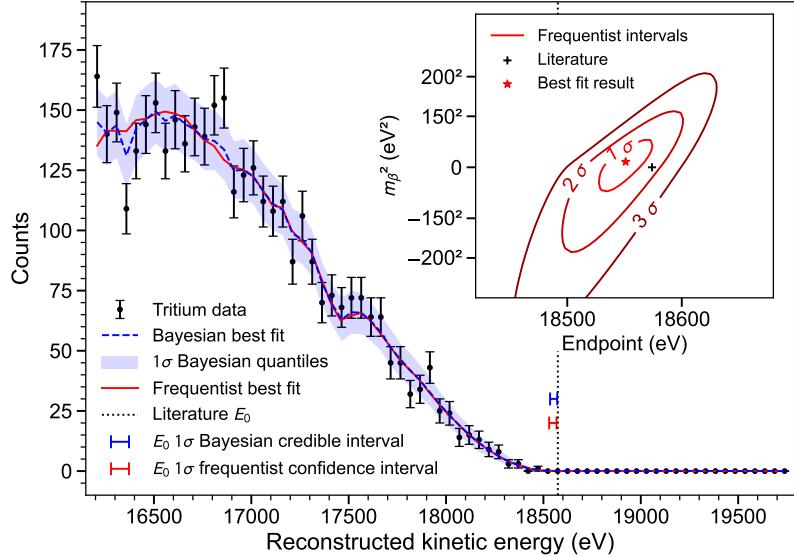


Figure 3.14: The measured tritium spectrum from Phase II with Bayesian and frequentist fits.

used to study the detection efficiency of the Phase II apparatus. Variations in the detection efficiency as a function of frequency directly influences the measured shape of the continuous tritium spectrum, which can lead to errors in the neutrino mass estimate if not modeled appropriately. Using the field shifting solenoid the cyclotron frequency of the krypton 17.83 keV line was shifted across the full frequency range of the tritium spectrum data (see Figure 3.13). Variations in the deep trap krypton spectrum shape can be used to infer the detection efficiency as a function of frequency and correct for this affect in the tritium measurements.

### 1157 Tritium Spectrum and Neutrino Mass Results

1158 The tritium measurement campaign resulted in the collection of 82 days of detector  
 1159 live time during which 3770 total tritium events were detected. The track and event  
 1160 reconstruction analysis extracted the starting frequencies of these tritium events, which  
 1161 were used to build a frequency spectrum of tritium beta-decays. The resulting frequency  
 1162 spectrum was then converted to an energy spectrum using the information gleaned from  
 1163 the krypton measurement campaign to obtain the tritium beta-decay spectrum (see  
 1164 Figure 3.14).

1165 CRES is inherently a very low background technique with the dominant source of  
 1166 noise being random RF fluctuations. Monte Carlo simulations backed validated using

1167 measurements of the RF noise background were used to set track and event cuts to  
1168 guarantee that zero false events would occur over the duration of the experiment with  
1169 90% confidence. Notably, the measured spectrum has zero events beyond the tritium  
1170 spectrum endpoint, which allows us to constrain the background rate in the Phase II  
1171 apparatus to less than  $3 \times 10^{-10}$  counts/ev/s. Achieving a low background is critical for  
1172 future neutrino mass experiments that seek to measure the neutrino mass with less than  
1173 100 meV sensitivity.

1174 Bayesian and frequentist based fits to the measured tritium spectrum, incorporating  
1175 information gained about CRES systematics from the krypton measurements, were  
1176 performed to extract upper limits on the tritium beta-decay spectrum endpoint as well as  
1177 the neutrino mass. The estimated spectrum endpoints are  $18553^{+18}_{-19}$  eV for the Bayesian  
1178 analysis and  $18548^{+19}_{-19}$  eV for the frequentist analysis. The quoted uncertainties are  
1179  $1-\sigma$ , and both results are within  $2-\sigma$  of the literature endpoint value of 15574 eV. The  
1180 estimated neutrino mass for both results is consistent with  $m_\beta^2 = 0$ . The 90% confidence  
1181 upper limits for the Bayesian analysis is  $m_\beta < 155$  eV/c<sup>2</sup> and  $m_\beta < 152$  eV/c for the  
1182 frequentist analysis.

1183 Though the neutrino mass results from Phase II are not competitive with KATRIN  
1184 the experiment was a promising first step towards the development of more precise  
1185 neutrino mass measurements using CRES. The low-background and high-resolution  
1186 achievable with krypton measurements are promising features of the technique that were  
1187 demonstrated with the Phase II apparatus. As new technologies are developed to enable  
1188 CRES measurements in larger volume, many of the lessons learned from Phase II will  
1189 continue to influence the operation and design of future experiments.

## 1190 **3.4 Phase III R&D: Antenna Array CRES**

1191 The goal of Phase III in the Project 8 experimental program is to develop the technologies  
1192 and expertise required to build an experiment that uses CRES to measure the neutrino  
1193 mass with a target sensitivity of 40 meV. One of the key technologies is a method for  
1194 performing high resolution CRES measurements in a large volume, which allows one to  
1195 observe a sufficient quantity of tritium to measure the low-activity endpoint region of  
1196 the tritium spectrum.

### **3.4.1 The Basic Approach**

One possible approach, suggested in the original CRES publication [3], is to use many antennas to surround a volume of tritium gas in a magnetic field (see Figure 3.15). When a decay occurs the electron will begin to emit cyclotron radiation that can be collected by the array and used to perform CRES. Each antenna in the array collects only a small

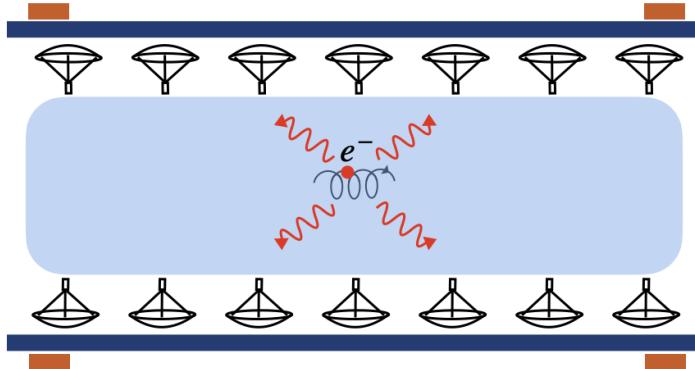


Figure 3.15: A cartoon illustration of the basics of the antenna array CRES technique.

fraction of the electron's signal power, which is less than 1 fW for a 18.6 keV kinetic energy electron in a 1 T magnetic field. Scaling to large volumes with the antenna array approach is accomplished by increasing the number of antennas in the array, which increases the volume under observation proportionally, so that a sufficient population of tritium atoms can be observed to measure the tritium spectrum endpoint shape.

Several features of the antenna array approach make it an attractive candidate technology for a large volume experiment. One example is the accurate position reconstruction made possible by the multichannel nature of the array. Using techniques like digital beamforming it is possible to estimate the radial and azimuthal positions of the electron in the magnetic trap with a precision significantly less than the size of the cyclotron wavelength. This capability allows one to perform event-by-event estimations of the magnetic field experienced by an electron, which is crucial to achieving high energy resolution with the CRES technique.

The easy availability of position information with the antennas array approach is potentially a unique advantage that provides significant flexibility in the magnetic field uniformity requirements compared to other proposed approaches to large volume CRES (see Chapter 6). Spatial discrimination using digital beamforming leads to pileup reduction, which helps to reduce the potential of background events caused by missing tracks or by incorrectly clustering a group of tracks into an event. Limits on the

background rate for a neutrino mass measurement with 40 meV sensitivity are stringent and the total activity of the tritium source for such an experiment is gigantic relative to the activity near the endpoint. Thus, pileup discrimination could be an important tool for a large scale CRES experiment.

Another beneficial quality of the antenna array approach is that the volume of the experiment can be scaled independent of frequency by simply adding more antennas to the array (see Figure 3.19). Resonant cavities, the proposed alternative large volume CRES technology, are ideally operated in magnetic fields that cause electrons to move with cyclotron frequencies near the fundamental cavity resonance, to avoid complex coupling of the electron to many cavity modes simultaneously. This leads to a coupling between the cavity volume and the magnetic field magnitude, which forces one to lower the magnetic field in order to increase the experiment scale. Whereas, for antenna arrays, in principle there is no physical limitation on the size of the antenna array that can be used at a particular magnetic field. However, the nature of scaling an antenna array based experiment leads to rapidly increasing cost and complexity due to the large number of antennas, amplifiers, and data streams that require substantial computer processing power to effectively analyze.

### 3.4.2 The FSCD: Free-space CRES Demonstrator

The complexity of the antenna array CRES technique requires the construction of a small scale demonstration experiment to develop an understanding of technique itself and relevant systematics. Without a demonstrator experiment it is not possible to sufficiently retire the technical risks associated with the full-scale experiment. Therefore, Phase III of the Project 8 experimental program is primarily focused on the development and operation of demonstrator experiments to inform the design of the Phase IV experiment.

The demonstrator experiment developed for antenna array CRES in Phase III is called the Free-space CRES Demonstrator or FSCD. The FSCD is intended as a demonstration of antenna array CRES, but is also a capable neutrino mass measurement experiment in its own right, with a target neutrino mass sensitivity of a few eV using a molecular tritium source.

#### Magnetic Field

The background magnetic field for the FSCD is provided by a hospital-grade MRI magnet (see Figure 3.16). The magnet produces a magnetic field of approximately 0.958 T, which

corresponds to a tritium spectrum endpoint frequency of approximately 25.86 GHz. The



Figure 3.16: An image of the MRI magnet installed in the Project 8 laboratory at the University of Washington, Seattle.

1253  
1254 magnet is installed in the Project 8 laboratory located at the University of Washington,  
1255 Seattle, and is shimmed to produce a uniform magnetic field with variations on the  
1256 ppm-level. Measurements of the magnetic field non-uniformities are performed using a  
1257 NMR probe and rotational gantry to capture measurements of the magnetic field around  
1258 an elliptical surface in the center of the MRI magnet. During the operation of the FSCD  
1259 an array of Hall or NMR magnetometers would be used to periodically measure the  
1260 magnetic field to monitor its time stability.

1261 Inside the field of the MRI magnet additional electromagnets would be installed that  
1262 provide the capability to shift the value of the background magnetic field and produce  
1263 a magnetic trap. Shifting the background magnetic field by a few  $\mu$  T lets one control  
1264 the cyclotron frequencies of electrons with a fixed kinetic energy, which is key to an  
1265 effective calibration of the FSCD. The preferred calibration method for the FSCD is  
1266 a mono-energetic electron gun that can inject electrons into the magnetic trap with a  
1267 known kinetic energy. In combination with the field shifting magnet, one can vary the  
1268 cyclotron frequencies of the electrons to measure the response of the antenna array as a  
1269 function of the radiation frequency and electron position. This procedure characterizes  
1270 the response of the antenna array and provides further information on magnetic field  
1271 uniformity, which important to achieving good energy resolution.

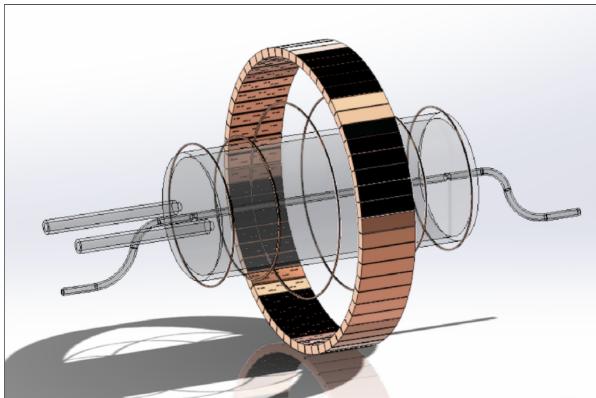
1272 The design of the magnetic trap is absolutely critical to the success of a CRES  
1273 experiment. The ideal shape is the perfect magnetic box, which has a flat bottom and

1274 step function walls. Any variation in the average magnetic field experienced by an  
1275 electron leads to changes in the cyclotron frequency that can make determining the true  
1276 starting kinetic energy more difficult. This includes changes in the magnetic field caused  
1277 by the walls of the magnetic trap as well as radial magnetic field variations.

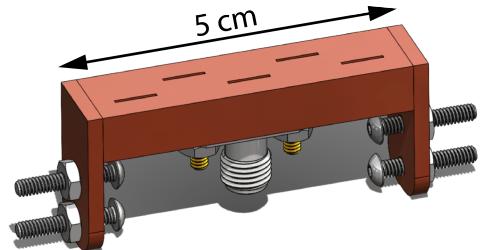
1278 The ideal box trap is completely uniform and has infinitely steep walls that cause  
1279 no change in the electron's cyclotron frequency as it is reflected from the trap wall,  
1280 however, such a trap cannot be made from any combination of magnetic coils since it  
1281 violates Maxwell's equations. One of the goals of magnetic trap design is to identify the  
1282 configuration of coils that produces a trap that approximates the perfect box trap as  
1283 closely as possible.

1284 **Antenna Array**

1285 The canonical antenna array design for CRES is a uniform cylindrical array of antennas  
1286 that surrounds the magnetic trap volume. Since the FSCD is a demonstrator experiment,  
1287 the antenna array design is the simplest form of the uniform cylindrical array, which is a  
single circular ring of antennas with a diameter of 20 cm (see Figure 3.17). Along this



(a)



(b)

Figure 3.17: (a) A model of the FSCD antenna array, magnetic trap, and tritium containment vessel design.(b) A more detailed model of a prototype design for the 5-slot waveguide antenna design.

1288  
1289 circle are sixty slotted waveguide antennas that fully populate the available space around  
1290 the array circumference. In order to maximize the power collected from each electron  
1291 it is optimal to cover as large a fraction of the solid angle around the magnetic trap as  
1292 possible.

1293 The distance between antennas around the circumference of the array is proportional

1294 to the wavelength of the cyclotron radiation. Therefore, maximizing the solid angle  
1295 coverage of the array, while minimizing channel count to keep the hardware and data  
1296 acquisition costs manageable, biases one towards smaller array diameters. Antenna  
1297 near-field effects limit the minimum diameter of the array for a given antenna design  
1298 since the radiation from electrons that are too close to the array cannot be detected  
1299 due to destructive interference caused by path-length differences from the electron to  
1300 different points on the antenna surface.

1301 Slotted waveguide antennas are used in the FSCD antenna array due to their high  
1302 efficiency and low loss, which comes from the lack of dielectric materials in the antenna  
1303 structure. Coupling to the waveguide can be performed with a coaxial cable connected  
1304 at the center or on either end of the waveguide. One of the drawbacks of waveguide  
1305 antennas is the large amount of space required to fit them inside the limited MRI magnet  
1306 volume. Alternative antenna designs, constructed from microstrip printed circuit boards  
1307 require significantly less space at the cost of slightly higher energy loss in the antenna  
1308 structure.

1309 The FSCD antenna design is a 5 cm long segment of WR-34 waveguide with 5 vertical  
1310 slots cut into the side. The distance between slots along the length of the waveguide is  
1311 a half wavelength for optimal power combination between the individual antenna slots.  
1312 Each slot is offset from the center of the antenna face a small distance in order to most  
1313 effectively couple the slot to waveguide modes inside the antenna.

1314 The passive power combination achieved by placing 5 slots in a single waveguide is a  
1315 compromise intended to reduce the cost and complexity of the antenna array system.  
1316 Each additional channel in the array requires it's own cryogenic amplifier and also increase  
1317 the required computer power to process the raw data collected by digitizing each channel.  
1318 Passive summation, achieved by combining antennas into arrays axially, reduces the array  
1319 channel count at the cost of losses from imperfect passive combination. Imperfect passive  
1320 combination is caused by effects such as re-radiation of energy from and destructive  
1321 interference between slots in the waveguide antenna.

1322 Interference and re-radiation eventually limit the achievable the axial extent of passive  
1323 power combination. The 5-slot designed developed for the FSCD is optimized to minimize  
1324 the impact of these losses while achieving the maximum amount of axial coverage with a  
1325 single ring of antennas. Scaling beyond the volume covered by a single ring of antennas is  
1326 achieved by stacking additional rings of antennas together to cover a larger trap volume  
1327 for a higher statistics measurement of the tritium spectrum endpoint region. A likely  
1328 scenario for the FSCD experiment involves a staged experiment approach, where first

1329 a series of measurements is performed using only a single ring of antennas followed by  
1330 experiments that add additional rings to the FSCD. The goal would be to first understand  
1331 the principles of antenna array CRES using the simplest possible experiment, before  
1332 attempting to scale the technique by expanding the antenna array size.

### 1333 **Tritium Source**

1334 While the primary purpose of the FSCD is as a technology demonstrator, it is unlikely  
1335 for the collaboration to gain the required confidence in the antenna array CRES tech-  
1336 nique to perform neutrino mass measurements at the 40 meV sensitivity level without  
1337 an intermediate scale measurement of the neutrino mass using antenna array CRES.  
1338 Therefore, the FSCD has an additional scientific goal of measuring the neutrino mass  
1339 with a rough sensitivity goal of a few eV. This level of precision is achievable using a  
1340 source of molecular tritium with a volume of approximately 1 L at a density comparable  
1341 to potential Phase IV scenarios.

1342 Unlike previous CRES experiments, where the tritium source could be co-located  
1343 with the receiving antenna inside a waveguide transmission line, the tritium source  
1344 in the FSCD is thermally isolated from the antenna array to avoid freeze-out of the  
1345 tritium molecules. The tiny radiation power emitted by electrons requires a system noise  
1346 temperature of  $\approx 10$  K or less, in order to detect events at a high enough efficiency to  
1347 reach the neutrino mass sensitivity goals of the experiment. Achieving a system noise of  
1348 10 K requires that the antenna array and amplifiers operate at cryogenic, liquid helium  
1349 temperatures of  $\approx 4$  K, which significantly lowers the vapor pressure of molecular tritium.  
1350 By keeping the molecular tritium isolated in an RF-transparent vessel the tritium gas can  
1351 be kept at a relatively warmer temperature in the range of 30 K to avoid the accumulation  
1352 of tritium on the experiment surfaces.

### 1353 **Data Acquisition and Reconstruction**

1354 A fundamental change in the data acquisition system for the FSCD is the shift from  
1355 single to multi-channel reconstruction. This transition results in a significant increase in  
1356 the data-generation rate, which is linearly related to the number of independent channels  
1357 in the array. The larger data volume coincides with an increased demand for computer  
1358 processing power based on the need for more precise signal reconstruction algorithms  
1359 driven by the FSCD and Phase IV sensitivity goals. Therefore, the data acquisition  
1360 system for the FSCD is likely to represent a significantly larger fraction of the experiment  
1361 cost and complexity than previous CRES experiments.

1362        Each antenna in the array is connected to a cryogenic amplifier and down-converted  
1363        from the 26 GHz CRES frequency using an IQ-mixer to reduce the size of the analysis  
1364        window in which the tritium spectrum is measured. Using an LO with a frequency of  
1365        approximately 25.80 GHz the antenna array signals can be digitized at a rate of 200 MHz,  
1366        which is sufficient bandwidth to resolve the complete sideband spectrum produced by  
1367        axial oscillations of electrons in the FSCD magnetic trap.

1368        Direct storage of the raw FSCD antenna array data is undesirable, since the estimated  
1369        amount of raw data generated is  $O(1)$  exabyte per year. The management and storage  
1370        of such a large dataset is infeasible for a demonstrator experiment on the scale of the  
1371        FSCD and would represent a large fraction of the budget for a Phase IV scale antenna  
1372        array based CRES experiment. Therefore, a sub-goal of the FSCD experiment is the  
1373        development of real-time reconstruction methods that could reduce the raw data volume  
1374        by detecting and reconstructing CRES events in real-time. The ultimate goal would be  
1375        a complete real-time reconstruction pipeline that takes raw voltages samples from the  
1376        antenna array and returns estimates for the starting kinetic energies of CRES events in  
1377        the data.

1378        The feasibility of a real-time reconstruction pipeline rests on the development of  
1379        computationally efficient algorithms that can be implemented without the need for  
1380        enormous computing resources. One challenge with the antenna array approach is that  
1381        the small radiation power of a single electron is distributed between each channel in  
1382        the array, such that reconstruction using only the information in a single channel is not  
1383        possible. Therefore, the simply performing the initial step in reconstruction — signal  
1384        detection — requires orders of magnitude more computational power than previous CRES  
1385        experiments. This operation will then be followed by other, potentially more expensive,  
1386        reconstruction steps that are required in order to determine the kinetic energy of the  
1387        electron.

## 1388        **3.5 Pilot-scale Experiments**

### 1389        **3.5.1 Choice of Frequency**

1390        The optimal CRES frequency for Project 8 is that which can reach our target sensitivity  
1391        of 40 meV, while minimizing the cost and complexity of the overall experiment. The  
1392        magnitude of the background magnetic field determines the cyclotron frequency, which  
1393        affects the entirety of the CRES detection system design, specifying the operating

1394 frequency of the CRES experiments is one of the first steps towards developing a full  
1395 design.

1396 **Scaling Laws**

1397 The Phase I and II experiments utilized a background magnetic field of 0.959 T provided  
1398 by an NMR magnet. This magnetic field was selected primarily for convenience, however,  
1399 the cyclotron frequencies for electrons near the tritium endpoint in a 0.959 T field ranges  
1400 from 25 to 26 GHz, which is within the standard RF Ka-band. Therefore, microwave  
1401 electronics specialized for these frequencies are easily obtainable for relatively low cost.  
1402 The operating frequency for the large-scale experiments must be selected in a more  
1403 rigorous manner due to the increased scale and complexity of the systems as well as the  
1404 requirements of the 40 meV neutrino mass science goal.

1405 There is a bias towards lower frequencies in a large-volume experiment, due to the  
1406 direct relationship between wavelength and the physical size of the compatible RF  
1407 components like antennas and cavities. With a longer wavelength more volume can  
1408 be surrounded by an array with fewer antennas, which reduces hardware and data-  
1409 processing costs. Additionally, the size of a cavity experiment is directly proportional  
1410 to the wavelength since this sets the physical dimensions of the cavity. Furthermore,  
1411 it is easier to engineer a magnet that provides a uniform magnetic field across several  
1412 cubic-meters of space at lower magnetic fields, which provides advantages in terms of  
1413 cost-reduction and field uniformity.

1414 A concern with lower magnetic fields and frequencies is the scaling of the Larmour  
1415 power equation, which is proportional to the square of the frequency. Naively, one would  
1416 predict that the SNR would decrease with lower fields, however, two additional scaling  
1417 laws that affect the noise power also come into play. Noise power is directly proportional  
1418 to the required bandwidth, which decreases linearly with the magnetic field. Furthermore,  
1419 at lower frequencies it is possible to purchase amplifiers with lower noise temperatures  
1420 until approximately 300 MHz at which point this relationship tends to flatten. Therefore,  
1421 it is expected that the SNR remains approximately constant as the frequency decreases.

1422 The SNR directly impacts the overall efficiency of the experiment through its effects  
1423 on signal detection and energy resolution. Thus, the expectation that SNR remains the  
1424 same at lower frequencies clearly biases large-scale experiments in this direction. One  
1425 drawback of lower magnetic fields is the increased influence of external magnetic fields  
1426 on the experiment. This includes magnetic fields from the building materials as well as  
1427 variations in the earth's magnetic field. To deal with these affects a suitable magnetic

<sup>1428</sup> field correction system will need to be devised, which includes constant monitoring of  
<sup>1429</sup> external fields.

### <sup>1430</sup> Atomic Tritium Considerations

<sup>1431</sup> The pilot-scale experiments will be the first Project 8 experiments to combine CRES with  
<sup>1432</sup> atomic tritium, therefore, the optimal frequency should take into account the affect of the background magnetic field on the atom trap. The primary influence of the background

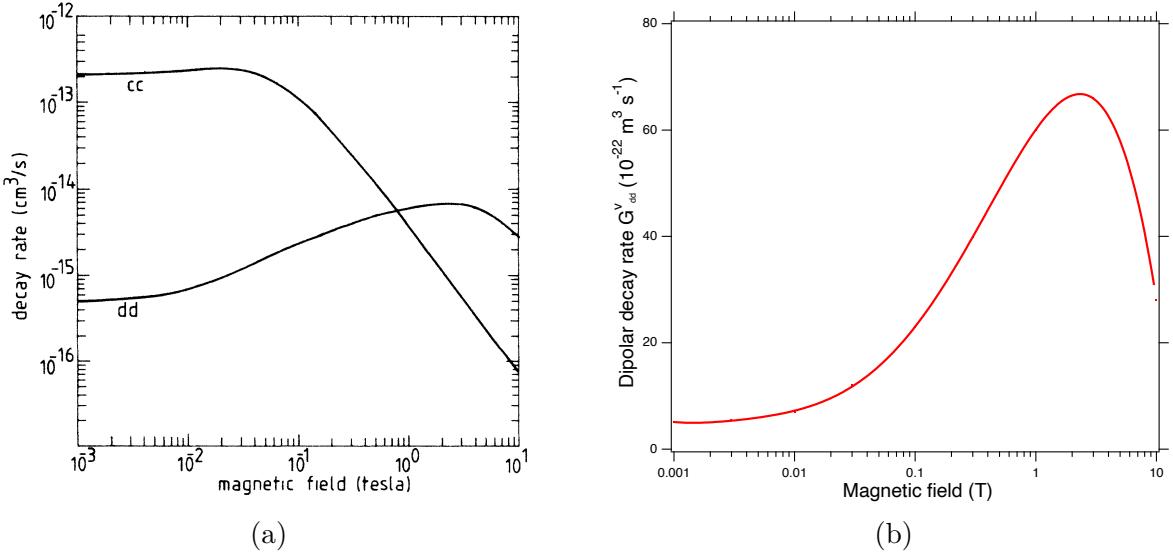


Figure 3.18: (a) A plot of the decay rate for the two-body dipolar spin exchange interaction for cc and dd state. (b) A plot of the decay rate of the dipolar spin exchange interaction for d+d states as a function of magnetic field magnitude. Lowering the magnetic field is key for reducing the losses from this interaction.

<sup>1433</sup>

<sup>1434</sup> field magnitude is through the rate of dipolar spin-flips caused by a spin exchange  
<sup>1435</sup> interaction between trapped atoms [21].

<sup>1436</sup> Atomic tritium is a simple quantum system with a hyperfine structure given by the  
<sup>1437</sup> addition of the nuclear and atomic spins. The addition of two spins leads to a hyperfine  
<sup>1438</sup> structure with four states in the  $(m_s, m_I)$  basis [22]. The states with atomic spins directed  
<sup>1439</sup> anti-parallel to the magnetic field have  $m_s = -1/2$  and are labeled as the a and b states.  
<sup>1440</sup> The a and b states are colloquially known as high-field seeking states, since their energy is  
<sup>1441</sup> minimized when in regions of higher magnetic field. This leads to losses in the magnetic  
<sup>1442</sup> trap as these atoms are drawn to higher fields away from the trap center. Alternatively,  
<sup>1443</sup> the c and d states, with atomic spin  $m_s = +1/2$ , minimize their energy in low magnetic  
<sup>1444</sup> fields because of the parallel alignment between spin and the magnetic field. Therefore,

1445 these low-field seeking states tend to stay trapped significantly longer than the high-field  
1446 seeking states.

1447 It would be advantageous to prepare tritium atoms in purely c and d states before  
1448 trapping, however, even in this case losses still occur due to dipolar interactions between  
1449 pairs of c and d states leading to flipped atomic spins and subsequent losses from high-field  
1450 seeking atoms. The rate of these interactions depends on the magnitude of the background  
1451 magnetic field and is maximal for dd interactions around 1 T (see Figure 3.18). The rate  
1452 of losses from these interactions at 1 T requires atomic tritium production at a rate two  
1453 orders of magnitude larger than at 0.1 T, thus, requirements on the whole atomic tritium  
1454 system are significantly relaxed at lower magnetic fields, which provides an additional  
1455 argument for transitioning to lower frequencies with the pilot-scale experiments.

1456 **3.5.2 Pilot-scale Experiment Concepts**

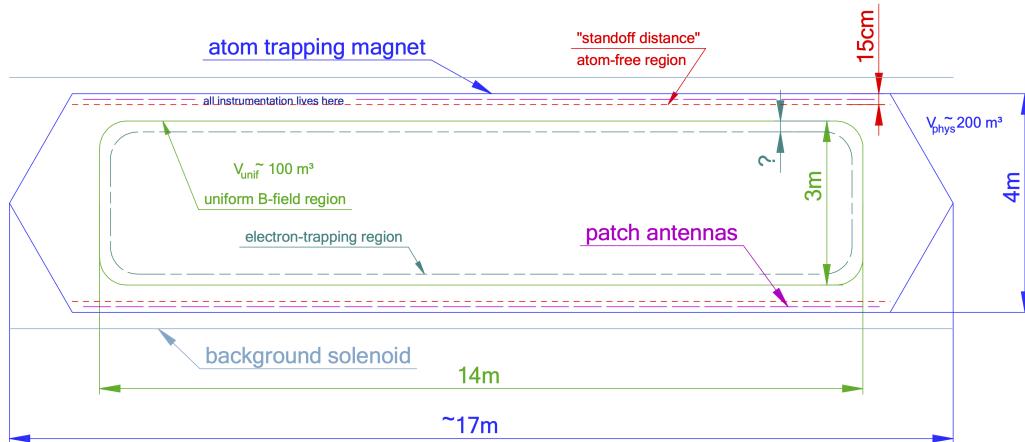


Figure 3.19: A conceptual sketch of a large-volume antenna array based CRES experiment to measure the neutrino mass.

1457 While the pilot-scale experiments are still in the early stages, enough is known to  
1458 sketch the general features of these experiments at the conceptual level.

1459 **Pilot-scale Antenna Array CRES Experiment Concept**

1460 A conceptual design for an antenna-based CRES experiment is shown in Figure 3.19.  
1461 A large solenoid magnet provides a uniform background magnetic field less than 0.1 T  
1462 in magnitude. Inside this region is the atom trapping magnet that generates a high  
1463 magnetic field at the walls, which decays exponentially towards the central region. Known

<sup>1464</sup> magnet designs that produce suitable atom trapping fields include Ioffe-Prichard traps,  
<sup>1465</sup> which use conducting coils, as well as a Halbach array made from permanent magnets.  
<sup>1466</sup> Either magnet choice produces a region of high magnetic fields, which excludes atoms  
<sup>1467</sup> and allows for the placement of antennas inside the experiment.

<sup>1468</sup> Inside this region an array of microstrip patch antennas is inserted to collect the  
<sup>1469</sup> cyclotron radiation without providing a surface for atomic tritium recombination. Due  
<sup>1470</sup> to the lower frequency of cyclotron radiation antennas of a larger size can be used,  
<sup>1471</sup> which lowers the total number of antennas required to observe the experiment volume.  
<sup>1472</sup> Because of this scaling, the lower frequency experiment uses a similar number of antennas  
<sup>1473</sup> compared to a much smaller demonstrator experiment with a 1 T magnetic field.

<sup>1474</sup> The atomic tritium beamline that supplies fresh tritium atoms to the experiment is  
<sup>1475</sup> not shown in the figure. The general configuration would matches the one shown for the  
<sup>1476</sup> pilot-scale cavity experiment (see Figure 3.20).

## <sup>1477</sup> Pilot-scale Cavity CRES Experiment Concept

<sup>1478</sup> The pilot-scale cavity experiment includes both an atomic tritium system and cavity  
<sup>1479</sup> CRES system. The atomic system consists of a thermal atom cracker located at the  
<sup>1480</sup> start of an evaporatively cooled atomic beamline. The atomic tritium system provides a  
<sup>1481</sup> supply of tritium atoms to the trap with temperatures on the order of a few mK. Atoms  
<sup>1482</sup> at this temperature can be trapped magneto-gravitationally, which is the reason for the  
<sup>1483</sup> vertical orientation of the cavity. At these low magnetic fields the trapping requirements  
<sup>1484</sup> for electrons and atoms differ enough such that it is advantageous to decouple the the  
<sup>1485</sup> trapping potentials to avoid radioactive heating of the tritium atoms from excess trapped  
<sup>1486</sup> electrons. Electron trapping is provided by a set of magnetic pinch coils at the top and  
<sup>1487</sup> bottom of the cavity and a multi-pole Ioffe or Halbach magnet serves to contain the  
<sup>1488</sup> atoms.

<sup>1489</sup> The cavity design for the pilot-scale experiment consists of a large cylindrical cavity  
<sup>1490</sup> with a TE011 resonance of 325 MHz. Such a cavity is truly enormous, with a diameter  
<sup>1491</sup> of approximately 1.2 m and a height of 11 m. When an electron is produced inside  
<sup>1492</sup> the cavity with a cyclotron frequency that matches the TE011 resonant frequency it's  
<sup>1493</sup> cyclotron orbit couples the electron to the TE011, which drives a resonance in the cavity.  
<sup>1494</sup> These resonant fields can be read-out using an appropriate cavity coupling mechanism  
<sup>1495</sup> located at the center of the cavity. For more information on the cavity approach to  
<sup>1496</sup> CRES see Chapter 6.

<sup>1497</sup> The bottom of the cavity has a cone termination to match the contour of the atom

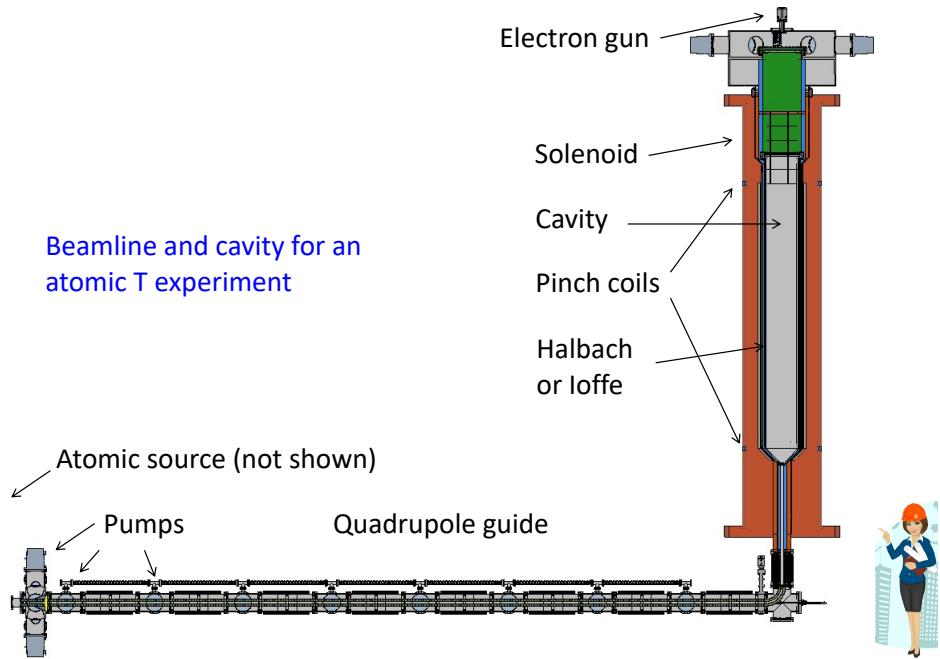


Figure 3.20: A conceptual sketch of a pilot-scale cavity CRES experiment with an atomic tritium beamline.

<sup>1498</sup> trapping magnet. This shape still allows for TE011 resonances with high internal Qs,  
<sup>1499</sup> which are required for good SNR in the cavity experiment. A small opening in the bottom  
<sup>1500</sup> of the cone serves as an entry point for the tritium atoms. To allow for calibration of  
<sup>1501</sup> the magnetic field inhomogeneities with an electron gun, the top of the cavity is left  
<sup>1502</sup> nearly completely open. Normally, this would drastically lower the Q-factor of the TE011  
<sup>1503</sup> mode, but a specially configured coaxial partition is inserted at the top. This termination  
<sup>1504</sup> scheme is designed to act as a perfect short for the TE011 mode since the circular shape  
<sup>1505</sup> of the partition matches the electric field boundary conditions for the TE011 mode.  
<sup>1506</sup> Simulations with HFSS have confirmed that this design results in a high quality TE011  
<sup>1507</sup> resonance despite the nearly completely open end.

## <sup>1508</sup> 3.6 Phase IV

<sup>1509</sup> The baseline CRES technology being pursued by the Project 8 collaboration are resonant  
<sup>1510</sup> cavities, which, due to their geometric properties, simple CRES signal structure, and low  
<sup>1511</sup> channel count, appear to be the better option for Phase IV. The current knowledge of the  
<sup>1512</sup> antenna array CRES approach reveals no technical obstacles that would preclude it as a  
<sup>1513</sup> baseline technology for Phase IV though it would most certainly be significantly more

1514 expensive. Therefore, antenna arrays represent a fallback approach if resonant cavities  
1515 prove infeasible.

1516 The sensitivity of the pilot-scale atomic tritium experiment is estimated to be on  
1517 the order of 0.1 eV, which means that increasing the sensitivity to reach the Phase IV  
1518 goal will require an even larger experiment. Because of the direct coupling between the  
1519 RF characteristics of a cavity and its geometry, the baseline plan is to build multiple  
1520 copies of the pilot-scale experiment (see Figure 3.21) to obtain the required amount of  
1521 volume rather than increase the size of the cavity beyond the pilot-scale. The built-in  
1522 redundancy of this approach is useful in the sense that the experiment has no single  
1523 point of failure, additionally, building several copies of the a pilot-scale experiment will  
1524 minimize new engineering and design effort.

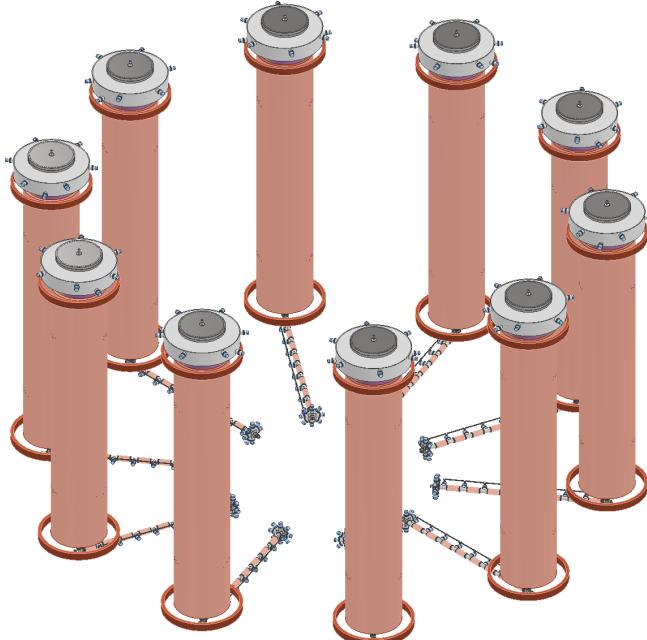


Figure 3.21: An illustration of a possible arrangement of ten pilot-scale cavity experiments for Phase IV. The experiments are arranged in a circle with an approximate diameter of 50 meters. Each atomic beamline connected to the bottom of each cavity is approximately 10 m in length. The cavities themselves are designed to operate at 325 MHz and are approximately 11 m tall. The circular arrangement of cavities has some advantages when it comes to cancellation of fringe fields from neighboring magnets, which is important due to the small magnetic field magnitudes consistent with these CRES frequencies. The advantage of ten independent atomic sources and cavities is that there is no single point of failure for the experiment. If an experiment goes down for repairs the other nine may continue running. Figure courtesy of Michael Huehn at UW-Seattle.

# **Chapter 4**

## **Signal Reconstruction Techniques for Antenna Array CRES and the FSCD**

### **4.1 Introduction**

The transition from a waveguide CRES experiment to an antenna array CRES experiment introduces new challenges related to data acquisition, signal detection, and signal reconstruction caused by the multi-channel nature of the data. The development of signal reconstruction algorithms is crucial to the design of antenna array based experiments like the FSCD, because these algorithms directly influence the detection efficiency and energy resolution of the CRES experiment. In this Chapter I summarize my contributions to the development and analysis of signal reconstruction and detection algorithms for the FSCD experiment.

In Section 4.2 I discuss the primary tool for this work, which is the Locust simulations package developed by the Project 8 experiment. Locust is used to simulate CRES events in the detector. Locust uses Kassiopeia to calculate particle trajectory solutions for electrons in the magnetic trap. The trajectories are then used to calculate the response of the antenna array to the cyclotron radiation produced by the electron, which results in signals that can be used to analyze the performance of different signal reconstruction algorithms. More recently, Project 8 has developed CREsana, which is a new simulations package that takes a more analytical approach to CRES signal simulations for antenna arrays. Although CREsana signals were not used for the signal reconstruction algorithm development detailed here, we introduce the software as it plays a role in the antenna array measurements presented in Section 5.4.

In Section 4.3 I discuss the signal reconstruction and detection approaches analyzed for the FSCD experiment. In general there are two steps to signal reconstruction — detection and parameter estimation. With signal detection one is primarily concerned

1551 only with distinguishing between data that contains a signal versus data that contains only  
1552 noise, whereas, with parameter estimation one is interested in extracting the kinematic  
1553 parameters of the electron encoded in the cyclotron radiation signal shape. Due to  
1554 the low signal power of electrons near the spectrum endpoint in the FSCD experiment,  
1555 signal detection is a non-trivial problem. This is magnified by the need to maximize the  
1556 detection efficiency of the experiment in order to achieve the neutrino mass sensitivity  
1557 goals. My contributions to signal reconstruction analysis for the FSCD are focused on  
1558 this signal detection component of reconstruction.

1559 After the discussion of various signal detection approaches, in Section 4.4 I present a  
1560 more detailed analysis of the detection performance of three algorithms, which could be  
1561 used to signal detection in the FSCD. This section was originally prepared for publication  
1562 in JINST as a separate paper. The algorithms include a digital beamforming algorithm,  
1563 a matched filter algorithm, and a neural network algorithm, which I analyze in terms of  
1564 classification accuracy and estimated computational cost.

## 1565 **4.2 FSCD Simulations**

1566 Antenna array CRES and the FSCD requires a combination of different capabilities  
1567 not often found in a single simulation tool. First of all, accurate calculations of the  
1568 magneto-static fields produced by current-carrying coils are required in order to accurately  
1569 model the magnetic trap and background magnets. The resulting magnetic fields must  
1570 then be used to calculate the exact relativistic trajectory of electrons, which is required  
1571 in order to calculate the electro-magnetic (EM) fields produced by the acceleration of  
1572 the electron. Finally, the simulation has to model the interaction of the antenna and  
1573 RF receiver chain with these EM-fields in order to produce the simulated voltage signals  
1574 produced by the antenna array during the CRES event. At the time when Project 8 was  
1575 developing this simulation capability, no single available simulation tool was known to  
1576 adequately perform this suite of calculations, which prompted the development of custom  
1577 simulation framework to simulate the FSCD. This simulation framework includes custom  
1578 simulation tools developed by Project 8 as well as other open-source and proprietary  
1579 software developed by third-parties.

1580 **4.2.1 Kassiopeia**

1581 Kassiopeia<sup>1</sup> is a particle tracking and static EM-field solver developed by the KATRIN  
1582 collaboration for simulations of their spectrometer based on magnetic adiabatic collimation  
1583 with an electrostatic filter [23]. Due to the measurement technique employed by the  
1584 KATRIN collaboration, Kassiopeia is not designed to solve for the EM-fields produced by  
1585 electrons in magnetic fields. However, it does provide efficient solvers for static electric  
1586 and magnetic fields and charged particle trajectory solvers. Because of this, Project 8  
1587 has incorporated parts of Kassiopeia into its own simulation framework.

1588 **Magnetostatic Field Solutions**

1589 The solutions to the electric and magnetic fields generated by a static configuration of  
1590 charges and currents is given by Maxwell's equations in the limit where the time-dependent  
1591 terms go to zero. In their static form Maxwell's equations [12] are

$$\nabla \cdot \mathbf{E} = \frac{\rho}{\epsilon_0} \quad (4.1)$$

$$\nabla \times \mathbf{E} = 0 \quad (4.2)$$

$$\nabla \cdot \mathbf{B} = 0 \quad (4.3)$$

$$\nabla \times \mathbf{B} = \mu_0 \mathbf{J}, \quad (4.4)$$

1592 where we can see that the electric and magnetic fields are now completely decoupled  
1593 from each other. The solution for the magnetic field in this boundary value problem is  
1594 given by the Biot-Savart law

$$\mathbf{B}(\mathbf{r}) = \frac{\mu_0}{4\pi} \int dr' \frac{r'^3 \mathbf{J}(\mathbf{r}') \times (\mathbf{r} - \mathbf{r}')}{|\mathbf{r}' - \mathbf{r}|^3}, \quad (4.5)$$

1595 which Kassiopeia uses a variety of numeric integration techniques to solve for a user  
1596 defined current distribution.

1597 **Kassiopeia Simulation of the FSCD Magnetic Trap**

1598 The trap developed for the FSCD experiment utilizes six current carrying coils, which  
1599 surround a cylindrical tritium containment vessel (see Figure 4.1). Some critical aspects  
1600 of the trap design include the total trapping volume, the maximum trap depth, the

---

<sup>1</sup><https://github.com/KATRIN-Experiment/Kassiopeia>

steepness of the trap walls, as well as the radial and azimuthal uniformity of the magnetic fields.

The volume of the FSCD trap is a cylindrically shaped region with a radius of 5 cm and a length of 15 cm resulting in a roughly 1 L total trap volume. The trap volume is an important design feature, because it sets the volume of the experiment that is potentially usable for CRES measurements. Trapping a larger volume allows one to observe a larger number of tritium atoms, which increases the statistical power and sensitivity of the neutrino mass measurement. Due to the cost of constructing magnets with large and uniform magnetic fields it is important that the trap use as much of the available volume as possible to limit the overall cost of the experiment.

Coil	Radius (mm)	Z Pos. (mm)	Current (A × Turns)
1	50.0	-92.3	750.0
2	50.1	-56.9	-220.3
3	68.5	-19.5	-250.0
4	68.5	19.5	-250.0
5	50.1	56.9	-220.3
6	50.0	92.3	750.0

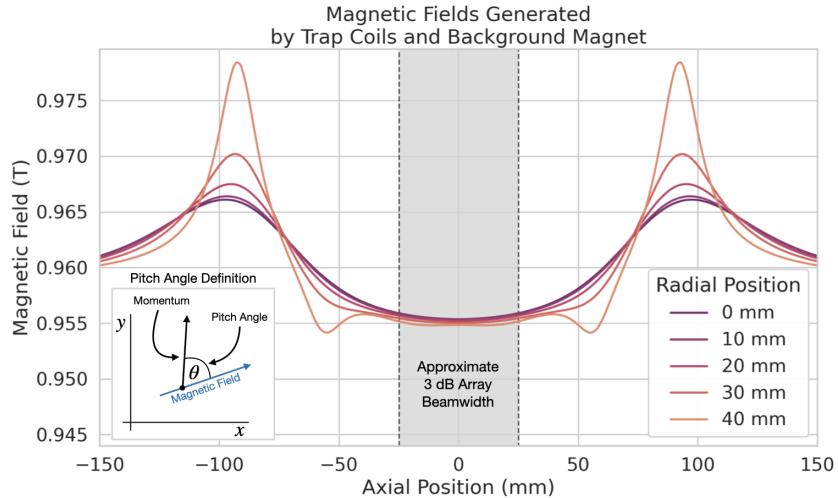
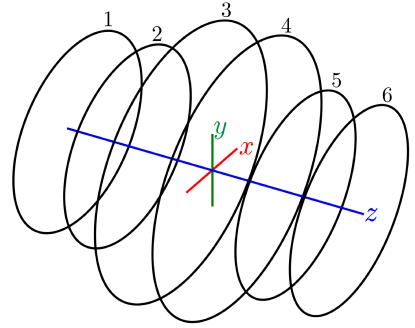


Figure 4.1: The geometry and parameters of the coils used to simulate the FSCD magnetic trap in Kassiopeia. Some axial profiles of the magnetic trap at different radial positions are shown to demonstrate the shape of the magnetic field and trap depth as a function of position. Calculation of the magnetic field profiles was graciously done by René Reimann.

The depth of the FSCD trap is approximately 10 mT when measured along the central axis, which is sufficient to trap electrons with pitch angles as small as  $84^\circ$ . The

trap depth factors into the efficiency of the experiment by directly controlling the range of electron pitch angles that can be trapped. If a higher fraction of pitch angles are trapped then, in principle, more decay events can be observed. However, the signals from electrons with small pitch angles are typically significantly harder to detect than larger pitch angles when using an antenna array, which increases the likelihood of not detecting the first track of the CRES event and harms the energy resolution of the experiment.

The steepness of the trap walls as well as any non-uniformities in the magnetic field contribute to the total energy resolution of the CRES measurement by causing uncertainty in the relationship between an electron's kinetic energy and it's cyclotron frequency. When an electron is trapped, it oscillates back and forth along the trap z-axis (see Figure 4.1) unless it is produced with a pitch angle of exactly 90° [24]. As the electron is reflected from the trap walls it experiences a change in the total magnetic field, which causes a modulation in the cyclotron frequency. This change in magnetic field from the trap introduces a correlation between the pitch angle and kinetic energy parameters of the electron that can reduce energy resolution. In order to mitigate this effect it is important to make the trap walls as steep as possible.

## Particle Trajectory Solutions

The magnetic fields solved by direct integration of the electron's current density can be used by Kassiopeia to solve for the trajectory of electrons based on user specified initial conditions. Various distributions are available within Kassiopeia that can be sampled in order to replicate realistic event statistics, including uniform, Gaussian, and Lorentzian among others. In general, an electron has six kinematic parameters that define its trajectory, which are the three-dimensional coordinates of the initial position and the three components of the electron's momentum vector. However, when simulating CRES events it is more common to parameterize the electron's trajectory in terms of it's initial position, the kinetic energy, the pitch angle, and the initial direction of the component of the electron's momentum perpendicular to the magnetic field. This parameterization is completely equivalent to specify each component of the electrons initial position and momentum vectors.

From the initial parameters of the electron and the magnetic field, Kassiopeia solves for the trajectory of the electron. The direct approach proceeds by solving the motion of the electron using the Lorentz force equation, which takes the form of a set of differential

1645 equations

$$\frac{d\mathbf{r}}{dt} = \frac{\mathbf{p}}{\gamma m} \quad (4.6)$$

$$\frac{d\mathbf{p}}{dt} = e(\mathbf{E} + \frac{\mathbf{p} \times \mathbf{B}}{\gamma m}), \quad (4.7)$$

1646 where  $\mathbf{r}$  is the position of the electron,  $\mathbf{p}$  is the electron's momentum,  $e$  is the charge of  
1647 the electron,  $m$  is the electron's mass, and  $\gamma$  is the relativistic Lorentz term. To account  
1648 for kinetic energy losses from radiation Kassiopeia includes an additional term in the  
1649 momentum differential equation, which calculates the change in the electron's momentum  
1650 induced by synchrotron radiation. Kassiopeia solves this pair of differential equations  
1651 using numerical integration, however, the exact trajectory can be computationally  
1652 intensive to solve. If the adiabatic approximation can be applied, then Kassiopeia can  
1653 make use of a simpler set of equations that can be more readily solved numerically.

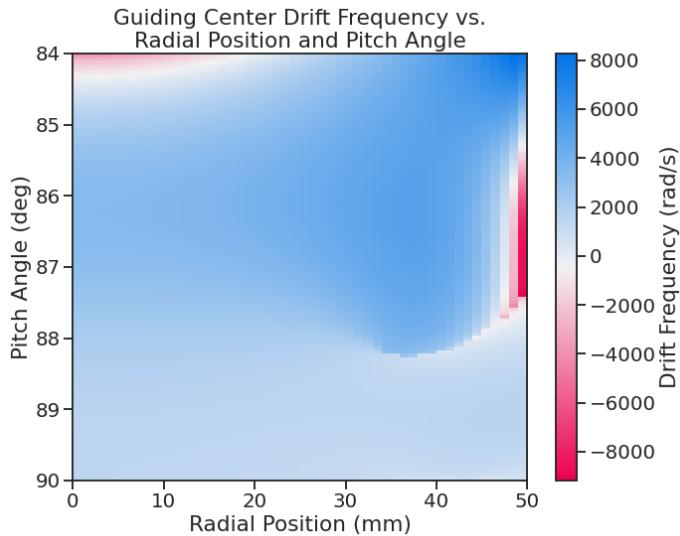


Figure 4.2: A map of the average  $\nabla B$ -drift frequency for electrons trapped in the prototype FSCD trap shown in Figure 4.1. Negative drift frequencies indicate electrons that are drifting opposite to the standard direction, which means that they are close to escaping the magnetic trap.

1654 Even though Kassiopeia is not directly capable of simulating the cyclotron radiation,  
1655 it is still an invaluable CRES simulation tool, due to the accurate trajectory solutions  
1656 for electrons in magnetic traps. With Kassiopeia it is possible to test the efficiency of a  
1657 particular trap design and analyze features of the electron trajectories that are important  
1658 to the position, track, and event reconstruction algorithms (see Section 4.3). One example

of this for the FSCD is the analysis of the average  $\nabla B$ -drift frequency as a function of  
 the electrons radial position and pitch angle in the magnetic trap (see Figure 4.2). Radial  
 gradients in the trap cause the guiding center of the electron to drift around the center of  
 the magnetic trap with an average frequency on the order of  $10^3$  rad/s. This frequency,  
 while slow compared to the length of a typical CRES time-slice, is large enough to cause  
 a significant loss in efficiency of certain signal reconstruction algorithms. Therefore, it is  
 important to model the drift of the electron in the reconstruction algorithm in order to  
 mitigate the effects of this motion on the reconstruction.

### 4.2.2 Locust

The Locust<sup>2</sup> software package [25] is the primary simulation tool developed and used  
 by the Project 8 collaboration for CRES experiments. Locust simulates the responses  
 of antennas and receiver electronics chain to rapidly time-varying electric fields using  
 a flexible approach that allows one to choose from a variety of electric field sources  
 and antennas. Similarly, one can simulate the receiver chain using a series of modular  
 generators that include standard signal processing operations such as down-mixing and  
 fast Fourier transforms (FFT). Since the primary focus of this chapter is the application  
 of Locust to analyses of the FSCD, we shall describe only the most relevant aspects of  
 the software rather than provide a comprehensive description.

### Cyclotron Radiation Field Solutions

Simulating CRES events in the FSCD requires that we calculate the electric fields  
 produced by the acceleration of the electron. In the general case, this can be a complicated  
 question to answer, due to back-reaction forces on the electron from its own electric fields  
 that occur when the electron is surrounded by conductive material such as a waveguide  
 or cavity. However, in the case of the FSCD it is possible to ignore such effects and  
 approximate the electron as radiating into a free-space environment.

The equations that describe the electromagnetic fields from a relativistic moving  
 point particle are the Liénard-Wiechert field equations [26, 27], which are obtained by  
 differentiating the Liénard-Wiechert potentials. In their full form the Liénard-Wiechert  
 field equations are

$$\mathbf{E} = e \left[ \frac{\hat{n} - \boldsymbol{\beta}}{\gamma^2(1 - \boldsymbol{\beta} \cdot \hat{n})^3 |\mathbf{R}|^2} \right]_{t_r} + \frac{e}{c} \left[ \frac{\hat{n} \times [(\hat{n} - \boldsymbol{\beta}) \times \dot{\boldsymbol{\beta}}]}{(1 - \boldsymbol{\beta} \cdot \hat{n})^3 |\mathbf{R}|} \right]_{t_r} \quad (4.8)$$

---

<sup>2</sup>[https://github.com/project8/locust\\_mc/tree/master](https://github.com/project8/locust_mc/tree/master)

$$\mathbf{B} = [\hat{n} \times \mathbf{E}]_{t_r}, \quad (4.9)$$

where  $e$  is the charge of the particle,  $\hat{n}$  is the unit vector pointing from the particle to the position where the fields are calculated,  $\beta$  and  $\dot{\beta}$  are the velocity and acceleration of the particle divided by the speed of light ( $c$ ),  $\mathbf{R}$  is the distance from the particle to the field calculation position, and  $\gamma$  is the relativistic Lorentz term. The subscript  $t_r$  indicates that the equations must be evaluated at the retarded time so that the time-delay from the travel time of the electromagnetic radiation is correctly accounted for.

The only required input to calculate the electric field at the position of an FSCD antenna is the velocity and acceleration of the electron, which can be obtained from Kassiopeia simulations. Therefore, when simulating a CRES event Locust first runs a Kassiopeia simulation of the electron and calculates the electric field incident on the antenna. The only difficulty with this approach is the determination of the retarded time. The retarded time corresponds to the time that a photon, which has just arrived at an antenna at the space-time position  $(t, \mathbf{r})$ , was actually emitted by the electron at the space-time position of  $(t_r, \mathbf{r}_e(t_r))$ . Defined in this way, finding the retarded time requires solving

$$c(t - t_r) = |\mathbf{r} - \mathbf{r}_e(t_r)|, \quad (4.10)$$

where the distance traveled by the photon between the measurement and retarded times is equal to the distance between the antenna and the electron at the retarded time. Locust solves Equation 4.10 using a built-in root finding algorithm to find the retarded time, and thus the electric field produced by the electron at the position of each antenna in the FSCD array.

## Antenna Response Modeling

With the electric field it is possible, in principle, to calculate the resulting voltages produced in the antenna. However, direct simulation of the antenna itself is computationally expensive since it would require the modeling of complex interactions of the electron's electric fields with charge carriers in the conductive elements of the antenna. Direct simulation of the antenna in Locust can be avoided by modeling the antenna response using the antenna factor, or antenna transfer function, approach. The antenna factor defines the voltage produced in the antenna terminal for an incident electric field [28],

$$A_F = \frac{V}{|\mathbf{E}|}, \quad (4.11)$$

1716 where  $V$  is the voltage and  $|\mathbf{E}|$  is the magnitude of the incident electric field. To obtain the  
 1717 antenna factor for the antennas developed for the FSCD Project 8 employs Ansys HFSS.  
 1718 HFSS is a commercially available finite element method electromagnetic solver widely  
 1719 used throughout the antenna engineering industry [29]. HFSS is capable of calculating  
 1720 the antenna factor and gain patterns for complex antenna designs and outputting the  
 1721 resulting quantities in the form of a text file that can be used as an input to the Locust  
 1722 simulation.

1723 The antenna factor defines the steady-state response of the antenna to electromagnetic  
 1724 plane waves and is a function of the frequency of the radiation. Therefore, in order to  
 1725 apply the transfer function for the calculation of the antenna voltage response in the  
 1726 time domain, Locust models the antenna as a linear time-invariant system [30]. In this  
 1727 formalism the response of the system to the driving force is given by

$$y[n] = h * x = \sum_k h[k]x[n - k], \quad (4.12)$$

1728 where  $y[n]$  is the discretely sampled response,  $x$  is the driving force stimulus, and  $h$  is  
 1729 the finite impulse response (FIR) filter. When applied to the FSCD array, this formalism  
 1730 calculates the voltage time-series produced in each antenna by convolving the electric  
 1731 field time-series with the antenna FIR filter, which is obtained by performing a inverse  
 1732 Fourier transform on the transfer function from HFSS.

### 1733 Radio-frequency Receiver and Signal Processing

1734 After obtaining the voltage time-series by computing the electron trajectory and antenna  
 1735 response, Locust simulates the signal processing associated with the radio-frequency  
 1736 receiver chain. The standard receiver chain used in Locust simulations of the FSCD  
 1737 attempts to mimic the operations that would actually occur in hardware (see Figure 4.3).

1738 Frequency down-conversion is used in the FSCD to reduce the digitization bandwidth  
 1739 required to read-out CRES data. According to the Nyquist sampling theorem [31], the  
 1740 minimal sampling rate that guarantees no information loss for a signal with a bandwidth  
 1741  $\Delta f$  is given by

$$f_{\text{Nyq}} = 2\Delta f. \quad (4.13)$$

1742 The total bandwidth of CRES signal frequencies from tritium beta-decay ranges from 0  
 1743 to 26 GHz in a 0.95 T magnetic field, therefore, direct digitization of CRES signals from  
 1744 the FSCD would require sampling frequencies greater than 50 GHz, which is infeasible for

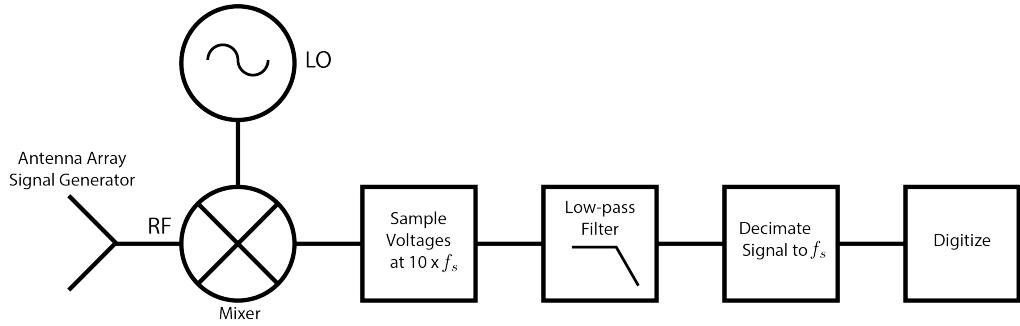


Figure 4.3: The receiver chain used by Locust when simulating CRES events in the FSCD.

1745 a real experiment. However, for the purposes of neutrino mass measurement we are only  
 1746 interested in measuring the shape of the spectrum in the last 100 eV, which corresponds  
 1747 to a frequency bandwidth of 5 MHz. Down-conversion is a technique for reducing the  
 1748 base frequencies of signals in a bandwidth given by  $[f_{\text{LO}}, f_{\text{LO}} + \Delta f]$  to the bandwidth  
 1749  $[0, \Delta f]$ , by performing the following multiplication

$$x(t) \rightarrow x(t)e^{-2\pi f_{\text{LO}} t}. \quad (4.14)$$

1750 In down-conversion the signal ( $x(t)$ ) is multiplied by a sinusoidal signal with frequency  
 1751  $f_{\text{LO}}$  to reduce the absolute frequencies of the signals in the bandwidth. In the FSCD this  
 1752 allows us to detect events in the last 100 eV of the tritium spectrum while sampling the  
 1753 data far below 50 GHz. The standard bandwidth used in the FSCD is 200 MHz, which  
 1754 allows for higher frequency resolution than the minimum sampling frequency for 100 eV  
 1755 of energy bandwidth.

1756 Trying to directly simulate down-conversion with a frequency multiplication in Locust  
 1757 would require the sampling of the electric fields at each antenna in the FSCD array with  
 1758 a period of  $\approx 20$  ps, which is extremely slow computationally. To avoid this Locust  
 1759 performs the down-conversion by intentionally under-sampling the electric fields with  
 1760 a frequency of 2 GHz. Sampling below the Nyquist limit causes the higher frequency  
 1761 components of the CRES signal to alias, however, Locust can remove these aliased  
 1762 frequency peaks using a combination of low-pass filtering and decimation to recreate  
 1763 frequency down-conversion. After filtering and decimation, Locust simulates digitization  
 1764 by an 8-bit digitizer at a sampling frequency of 200 MHz to recreate the conditions of  
 1765 the FSCD. The voltage offset and the digitizer range must be configured by the user  
 1766 based on the characteristics of the simulation.

1767 **Data**

1768 The output of Locust simulations for the FSCD primarily consists of two data files. The  
1769 first is the electron trajectory information calculated by Kassiopiea, which is output in  
1770 the form of a `.root` file [32]. This file contains important kinematic information about  
1771 the electron such as it's position and pitch angle as a function of time. The other file is  
1772 produced by Locust and it contains the digitized signals acquired from each antenna in  
1773 the FSCD array. The Locust output files conform to the Monarch specification developed  
1774 by Project 8, which is based on the commonly used HDF5 file format, and matches the  
1775 format of the files produced by the Project 8 data acquisition software. This makes it  
1776 possible to use the same data analysis code to analyze both simulated and real data.

1777 **4.2.3 CRESana**

1778 Locust is the primary simulation tool used by Project 8 in the development and simulation  
1779 of the FSCD. However, simulations of CRES events in larger antenna arrays ( $\geq 100$   
1780 antennas) using Locust can take several hours to complete, which is prohibitively long  
1781 when one is performing a sensitivity analysis for a large scale antenna experiment. One  
1782 of the reasons for Locust's slow operation is that the electric fields from the electron  
1783 must be solved numerically for each time-step for each of the antennas in the array.  
1784 These numerical solutions allow Locust to accurately simulate the electric fields from  
1785 arbitrarily complicated electron trajectories at the cost of more computations and slower  
1786 simulations. Therefore, an additional simulation tool that sacrifices some accuracy for  
1787 computational efficiency would be extremely useful simulations and sensitivity analyses  
1788 of larger antenna array experiments.

1789 To fill this need, Project has developed a new simulations package called CRESana<sup>3</sup>,  
1790 specifically designed to perform analytical simulations of antenna array based CRES  
1791 experiments. CRESana is not as flexible as Locust, but it provides a significant increase  
1792 in simulation speed. It does this by using well-justified analytical approximations of the  
1793 electrons motion in the magnetic field and the resulting electric fields from the electron's  
1794 acceleration. The electric fields and signals generated by CRESana are consistent with  
1795 theoretical calculations of the electron's radiation, and are test for accuracy using  
1796 well-known test-case simulations and consistency checks.

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3<https://github.com/MCFlowMace/CRESana>

## **4.3 Signal Detection and Reconstruction Techniques for Antenna Array CRES**

### **1799 Antenna Array CRES Signal Reconstruction**

**1800** A robust set of FSCD simulation tools are vital to the development of the analysis  
**1801** algorithms necessary for antenna array CRES to succeed. In order to perform CRES  
**1802** measurements using an antenna array, one must develop an algorithm that uses the  
**1803** multi-channel time-series obtained by digitizing the array to estimate the starting kinetic  
**1804** energies of electrons produced in the magnetic trap. This procedure consists of a multi-  
**1805** stage process of detecting a CRES signal then estimating the parameters of the electron  
**1806** that produced and is often referred to as simply CRES signal reconstruction.

**1807** Compared with the signal reconstruction approaches of the Phase I and II CRES  
**1808** experiments, antenna array CRES requires a significantly different approach to signal  
**1809** reconstruction. In Phase I and II, CRES was performed using a waveguide gas cell that  
**1810** could be directly connected to a waveguide transmission line. The transmission line  
**1811** efficiently transmits the cyclotron radiation along its length to an antenna at either end  
**1812** of the waveguide. However, with an antenna array the electron is essentially radiating  
**1813** into free-space, therefore, the cyclotron radiation power collected by the array is directly  
**1814** proportional to the solid angle surrounding the electron that is covered with antennas.  
**1815** Because it is not practical to fully surround the magnetic trap with antennas, some of the  
**1816** cyclotron radiation power that would have been collected by the waveguide escapes into  
**1817** free-space. Furthermore, the power that is collected by the antenna array is split between  
**1818** every channel in the antenna array, which significantly lowers the signal-to-noise ratio  
**1819** (SNR) of CRES signals in a single antenna channel compared to a waveguide apparatus.  
**1820** Therefore, a suite of completely new signal reconstruction techniques are needed in order  
**1821** to perform CRES in the FSCD.

**1822** Changes to the approach to CRES signal reconstruction are also motivated by the  
**1823** more ambitious scientific goals of the FSCD experiment. A measurement of the tritium  
**1824** beta-decay spectrum that is sensitive to neutrino masses as small as 40 meV requires that  
**1825** we measure the kinetic energies of individual electrons with a total energy broadening  
**1826** of 115 meV [33]. This resolution includes all sources of uncertainty in the electron's  
**1827** kinetic energy such as magnetic field inhomogeneities. This level of energy resolution is  
**1828** compatible only with an event-by-event signal reconstruction approach where the kinetic  
**1829** energies, pitch angles, and other parameters of the CRES events are estimated before

1830 constructing the beta-decay spectrum.

1831 The event-by-event approach is distinct from the analysis done for the Phase I and  
1832 Phase II experiments where only the starting cyclotron frequency of the event was  
1833 estimated by analyzing the tracks formed by the carrier frequency in the time-frequency  
1834 spectrogram. These frequencies were then combined into a frequency spectrogram, which  
1835 was converted to the beta-decay energy spectrum using an ensemble approach that  
1836 averaged over all other event parameters. The ensemble approach to signal reconstruction  
1837 results in poor energy resolution because other kinematic parameters such as pitch angle  
1838 change the cyclotron carrier frequency due to changes in the average magnetic field  
1839 experience by the electron, and it is therefore incompatible with the future goals of the  
1840 Project 8 collaboration.

#### 1841 Components of Reconstruction: Signal Detection and Parameter Estimation

1842 CRES signal reconstruction can be viewed as a two-step procedure consisting of signal  
1843 detection followed by parameter estimation. In the former, one is concerned with  
1844 identifying CRES signals in the data regardless of the signal parameters, whereas, in the  
1845 latter one operates under the assumption that a signal is present and then estimates its  
1846 parameters.

1847 More formally, signal detection is essentially a binary hypothesis test between the  
1848 signal and noise data classes and parameter estimation describes a procedure of fitting a  
1849 model to the observed data. While both of these processes are required for a complete  
1850 reconstruction (see Figure 4.4), the focus of my work and this chapter is on the signal  
1851 detection aspect of antenna array CRES signal reconstruction.

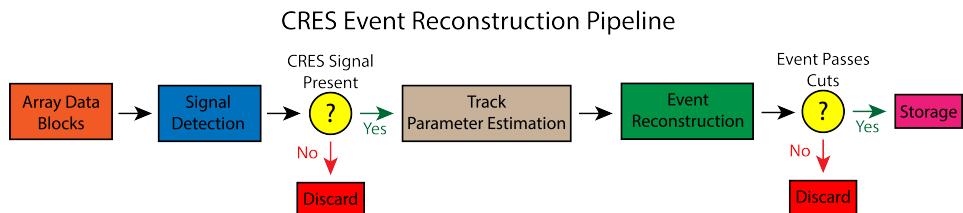


Figure 4.4: A high-level diagram depicting the process of CRES event reconstruction. The first step consists of identifying the presence of a signal in the data. This step is necessary to avoid the danger of performing a reconstruction of a false event, which would constitute a background contribution to the tritium spectrum measured by CRES.

1852 **Detection Theory**

1853 The problem of signal detection can be posed as a statistical hypothesis test [34]. For  
1854 CRES signals, which are essentially vectors with added white Gaussian noise (WGN),  
1855 one needs to choose between two hypotheses

$$\mathcal{H}_0 : \mathbf{y} = \boldsymbol{\nu} \quad (4.15)$$

$$\mathcal{H}_1 : \mathbf{y} = \mathbf{x} + \boldsymbol{\nu}, \quad (4.16)$$

1856 where  $\mathbf{y}$  is the CRES data vector,  $\boldsymbol{\nu}$  is a sample of WGN, and  $\mathbf{x}$  represents the CRES  
1857 signal. The hypothesis that the data contains only noise is labeled  $\mathcal{H}_0$  and the hypothesis  
1858 that the data contains a signal is labeled  $\mathcal{H}_1$ .

1859 For illustrative purposes one can examine the case where one the first sample of  
1860 data is used to distinguish between  $\mathcal{H}_0$  and  $\mathcal{H}_1$ . The value of the first data sample is  
1861 distributed according to two gaussian distributions corresponding to  $\mathcal{H}_0$  and  $\mathcal{H}_1$  (see  
1862 Figure 4.5). By setting a decision threshold on the value of this sample, one can choose  
1863 the correct hypothesis with a probability given by the areas underneath the probability  
1864 distribution curves. A true positive corresponds to correctly identifying that the data  
1865 contains signal, whereas, a true negative means that one has correctly identified the data  
1866 as noise. The rate at which the detector performs a true positive classification is given  
1867 by the green region underneath  $p(\mathbf{y}[0]; \mathcal{H}_0)$ , and the rate at which the detector performs  
1868 a true negative classification is given by the orange region underneath  $p(\mathbf{y}[0]; \mathcal{H}_1)$ . Two  
1869 types of misclassifications are possible. Either we declare noise data as signal, which is  
1870 call a false positive, or we declare signal data as noise, which is a false negative. Note  
1871 that it is only possible to trade off these two types of errors by tuning the detection  
1872 threshold. One cannot simultaneously reduce the rate of false positives without also  
1873 increasing the rate of false negatives.

1874 The approach taken with CRES signals is to fix the rate of false positives by setting  
1875 a minimum value for a detection threshold. The rate of false positives that is acceptable  
1876 at the detection stage depends upon the rate of background events compatible with the  
1877 sensitivity goals of the experiment. The ultimate goal of a neutrino mass measurement  
1878 with 40 meV sensitivity in general has strict requirements on the number of background  
1879 events, which requires a relatively high detection threshold to achieve. Consequently,  
1880 the ideal signal detection algorithm is the one that achieves the maximum rate of true  
1881 positives for a fixed rate of false positives, so that the detection efficiency of the experiment  
1882 is maximized and potential sources of background are kept to a minimum.

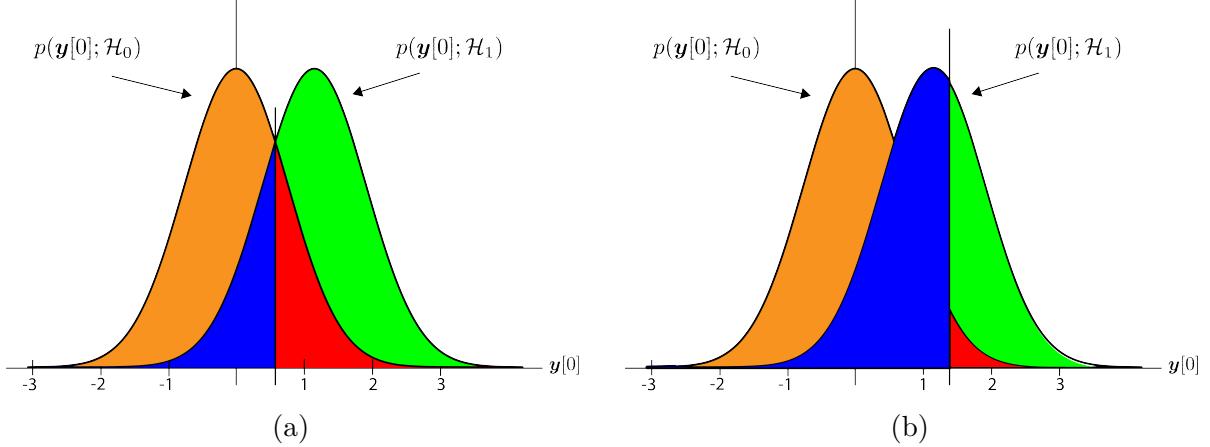


Figure 4.5: An illustration of two PDFs associated with a binary hypothesis test. The decision threshold is represented by the vertical line that partitions both distributions. The orange and red areas correspond to the true negative and false positive probabilities and the blue and green areas correspond to the false negative and true positive probabilities respectively. To decide between the two hypotheses we perform the likelihood ratio test specified by the Neyman-Pearson theorem. This approach achieves the highest true positive probability for a given false positive probability.

According to the Neyman-Pearson theorem [35], the statistical hypothesis test that maximizes the probability of detection for a fixed rate of false positives is the likelihood ratio test, which is formed by computing the ratio of the signal likelihood to the noise likelihood,

$$L(x) = \frac{P(\mathbf{y}; \mathcal{H}_1)}{P(\mathbf{y}; \mathcal{H}_0)} > \gamma. \quad (4.17)$$

Here, the likelihood of the hypotheses  $\mathcal{H}_0$  and  $\mathcal{H}_1$  are described by the probability distributions  $P(\mathbf{y}; \mathcal{H}_0)$  and  $P(\mathbf{y}; \mathcal{H}_1)$  respectively, and  $\gamma$  is the threshold for deciding  $\mathcal{H}_1$ . The decision threshold is determined by integrating  $P(\mathbf{y}; \mathcal{H}_0)$  such that

$$P_{\text{FP}} = \int_{\gamma}^{\infty} P(\tilde{\mathbf{y}}; \mathcal{H}_0) d\tilde{\mathbf{y}} = \alpha, \quad (4.18)$$

where  $\alpha$  is the desired false positive detection rate given by the red colored areas shown in Figure 4.5. The true positive detection rate is given by the similar integral

$$P_{\text{TP}} = \int_{\gamma}^{\infty} P(\tilde{\mathbf{y}}; \mathcal{H}_1) d\tilde{\mathbf{y}}, \quad (4.19)$$

which corresponds to the green areas in Figure 4.5.

Changing the decision threshold allows one to trade-off between  $P_{\text{TP}}$  and  $P_{\text{FP}}$  as

appropriate for the given situation. It is common to summarize the relationship between  $P_{\text{TP}}$  and  $P_{\text{FP}}$  using the receiver operating characteristic (ROC) curve, which is obtained by evaluating the true positive and false positive probabilities as a function of the decision threshold value (see Figure 4.6). The ROC curve provides a convenient way to compare

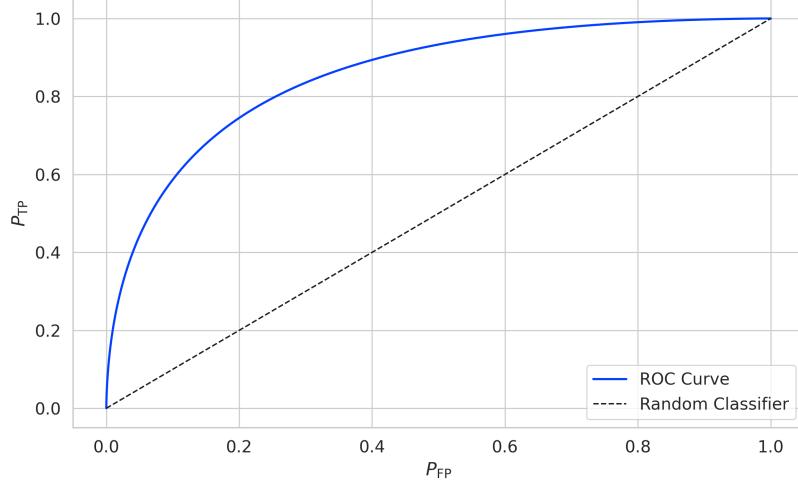


Figure 4.6: An example ROC curve formed by computing the  $P_{\text{FP}}$  and the  $P_{\text{TP}}$  for a given likelihood ratio test. As the decision threshold is increased  $P_{\text{FP}}$  decreases at the expense of a lower  $P_{\text{TP}}$ . The black dashed line indicates the lower bound ROC curve obtained by randomly deciding between  $\mathcal{H}_0$  and  $\mathcal{H}_1$ .

the performance of different signal detection algorithms. In general, a classifier with a higher the  $P_{\text{TP}}$  as a function of  $P_{\text{FP}}$  is desirable, which corresponds to a larger area underneath the respective ROC curve. A perfect classifier has an area underneath the curve of 1.0, however, such a classifier is almost never achievable in practice.

### 4.3.1 Digital Beamforming

#### Introduction to Beamforming

Beamforming refers to a suite of antenna array signal processing techniques that are designed to enhance the radiation or gain of the array in certain directions and suppress it in other direction [28]. Beamforming is of interest to Project 8 as a first level of signal reconstruction for the FSCD and other antenna array CRES experiments, which operates at the signal detection stage of reconstruction.

Beamforming is accomplished by performing a phased summation of the signals received by the antenna array. The beamforming phases are chosen such that the signals

1911 emitted by the array will constructively interfere at the point of interest (see Figure  
 1912 4.7). As a consequence of the principle of reciprocity [36], when the array is operating in  
 1913 receive mode, the signals emitted from a source at the same point will constructively  
 interfere when summed. The origin of the phase delays in beamforming is the path-

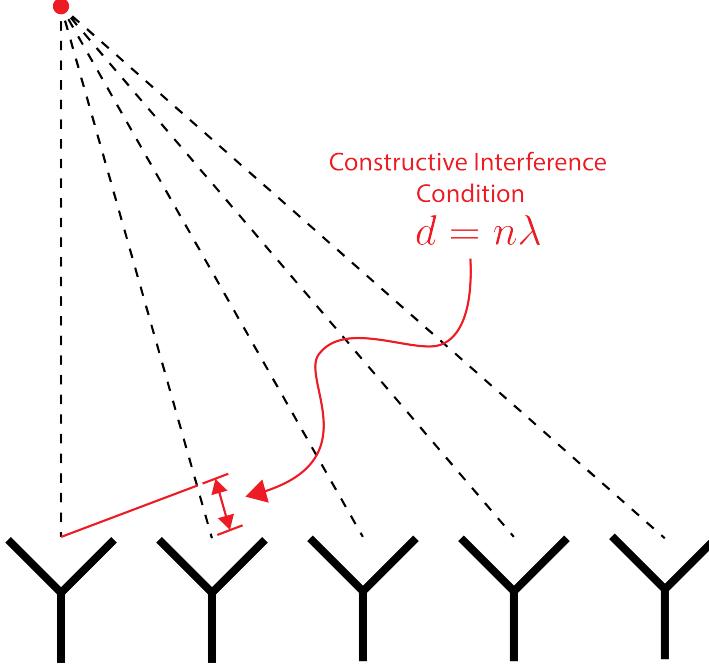


Figure 4.7: An illustration of the constructive interference condition which is the operating principle of digital beamforming using a uniform linear array as an example.

1914  
 1915 length difference to the beamforming point between different antennas in the array. The  
 1916 relationship between the phase delay and the path-length difference is given by the  
 1917 familiar equation

$$\phi = \frac{2\pi d}{\lambda}, \quad (4.20)$$

1918 where  $\phi$  is the phase delay,  $d$  is the path-length difference, and  $\lambda$  is the wavelength of  
 1919 the radiation. In practice, one chooses the values of  $d$  by specifying the beamforming  
 1920 positions of interest and then calculates the beamforming phases using Equation 4.20,  
 1921 which is guaranteed to follow the constructive interference condition shown in Figure 4.7.

1922 Beamforming can be neatly expressed mathematically using the vector equation

$$y[n] = \Phi^T[n] \mathbf{x}[n], \quad (4.21)$$

1923 where  $\mathbf{x}[n]$  is the array snapshot vector,  $\Phi[n]$  is a vector of beamforming shifts, and  
 1924  $y[n]$  is the resulting summed signal. The beamforming shifts consist of a set of complex

1925 numbers that contain the beamforming phase shift and an amplitude weighting factor,

$$\Phi[n] = [A_0[n]e^{-2\pi i \phi_0[n]}, A_1[n]e^{-2\pi i \phi_1[n]}, \dots, A_{N-1}[n]e^{-2\pi i \phi_{N-1}[n]}], \quad (4.22)$$

1926 where the set of magnitudes  $A_i[n]$  are amplitude weighting factors and  $\phi_i[n]$  are the phase  
1927 shifts from the path-length differences. The index  $i$  is used to denote the antenna channel  
1928 number. The amplitude weighting factor is the relative magnitude of the signal received  
1929 by a particular antenna to the other antennas in the array, such that the antennas that  
1930 receive signals with higher amplitude, due to being closer to the source, have more  
1931 weight in the beamforming summation. The input and outputs signals beamforming  
1932 are naturally expected to be functions of time as indicated by the index  $[n]$ , however, it  
1933 is also possible to use time dependent beamforming phases that shift the beamforming  
1934 position of the array over time.

1935 Digital beamforming is the type of beamforming algorithm of interest to Project 8 for  
1936 CRES. Specifically, digital beamforming means that the beamforming phases are applied  
1937 to the array signals in software rather than employing fixed beamforming phase shifts in  
1938 the receiver chain hardware. The advantage of digital beamforming is that for a given  
1939 series of array snapshots one can specify a large number of beamforming positions and  
1940 effectively search for electrons by performing the beamforming summation associated  
1941 with each point and applying a signal detection algorithm to identify the presence of a  
1942 CRES signal.

1943 One of the most attractive features of digital beamforming is the spatial filtering  
1944 effect, which is a direct consequence of the constructive interference condition used to  
1945 define the beamforming phases. Spatial filtering allows for signals from multiple electrons  
1946 at different positions in the trap to be effectively separated, because the constructive  
1947 interference condition will force the signals from electrons at positions different from the  
1948 beamforming position to cancel. This helps to reduce signal pile-up that could become  
1949 an issue for large scale CRES experiments using a dense tritium source.

1950 The digital beamforming positions can be specified with arbitrary densities limited  
1951 only by the available computational resources. This provides a very straight-forward way  
1952 to estimate the position of the electron in the trap by using a dense grid of beamforming  
1953 positions and maximizing the output power of the beamforming summation over this  
1954 grid. This natural approach to position reconstruction is attractive due the requirements  
1955 of an event-by-event signal reconstruction, which needs an accurate estimation of the  
1956 exact magnetic field experienced by the electron in order to correctly estimate it's kinetic

1957 energy. Combined with an accurate map of the magnetic field inhomogeneities of the  
1958 trap obtained from calibrations, beamforming allows one to apply this magnetic field  
1959 correction with a spatial resolution that is a fraction of the cyclotron wavelength.

1960 **Laboratory Beamforming Demonstrations**

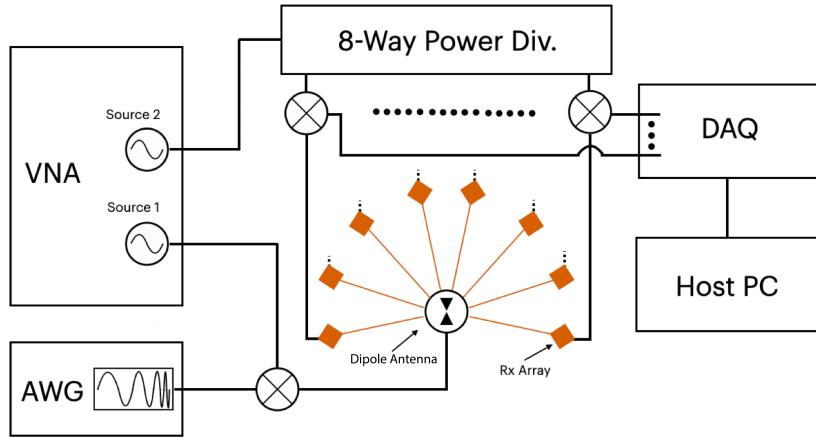


Figure 4.8: System level diagram of the laboratory setup used for beamforming demonstrations at Penn State. For more information on this system see Chapter 5. Signals near 26 GHz are fed to a dipole antenna using an arbitrary waveform generator (AWG) and vector network analyzer (VNA), which drive a mixer. The dipole radiation is collected by an array of antennas connected to the digitizer data acquisition (DAQ) system.

1961 As part of the development of antenna array CRES for the FSCD, an antenna  
1962 measurement setup was constructed at Penn State to serve as a testbed for antenna  
1963 prototypes and to perform laboratory validations of array simulations. This system  
1964 is discussed in more detail in Chapter 5. Early versions of the antenna measurement  
1965 system (see Figure 4.8 and Figure 4.9) were used to perform beamforming reconstruction  
1966 studies of a simple probe antenna to better understand the principles of beamforming  
1967 and confirm the estimated beamforming performance of Locust.

1968 Signals from an arbitrary waveform generator were up-converted to 26 GHz using a  
1969 mixer and a high-frequency source from a vector network analyzer and fed to the dipole  
1970 antenna through a balun. The radiation from the dipole antenna was received by an  
1971 array of horn antennas. The signals from the horn antennas were then down-converted  
1972 to baseband using a collection of mixers and an 8-way power divider. The signals were  
1973 then digitized and saved to a host computer for analysis.

1974 The data collected using the dipole and horn antenna array is reconstructed using the

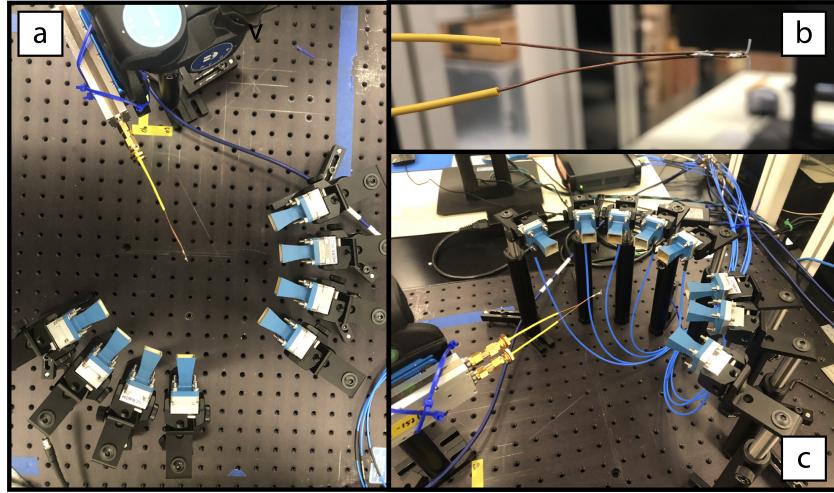


Figure 4.9: Photographs of the beamforming demonstration setup. In (a) I show a top-down view of the dipole antenna and the array of eight horn antennas. Manual repositioning of the horn antennas allows one to synthesize a full-circular antenna array. The dipole antenna is mounted on a camera tripod mount that allows for manual position tuning. (b) is a close up image of the dipole, which is manufactured from two segments of semi-rigid coaxial cable. (c) is another image of the dipole and array.

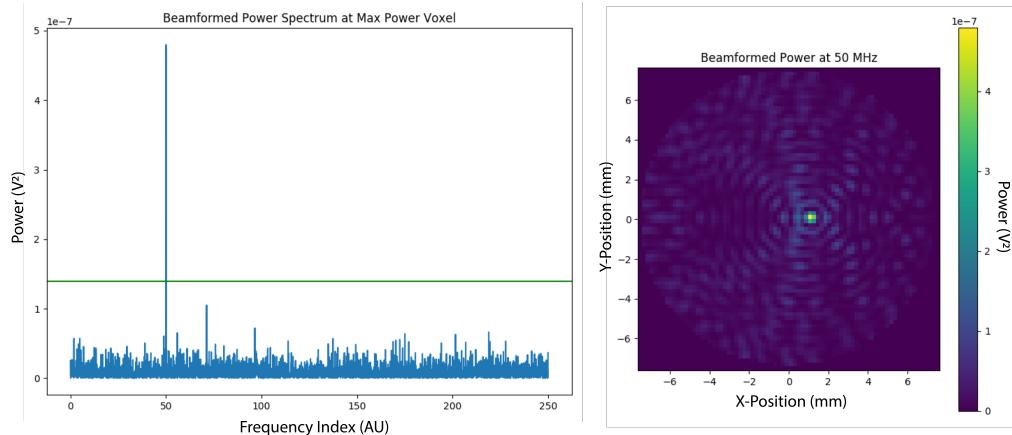


Figure 4.10: An example of digital beamforming reconstruction of a dipole antenna using a synthetic array of horn antennas. The beamforming image on the right is constructed by computing the time-averaged power of the summed signals for a two-dimensional grid of beamforming positions. In the image one can see a clear maximum that corresponds to the position of the dipole antenna. On the left I show the frequency spectrum of the time-series at the maximum power pixel. White gaussian noise is added to the signal to mimic a more realistic signal-to-noise-ratio. The signal emitted by the dipole is clearly visible as the high power peak in the frequency spectrum.

beamforming reconstruction approach specified in Section 4.3.1. A two-dimensional grid of xy-positions is defined and the beamforming phase shifts for each of these positions is calculated. The phased summation can be visualized by plotting the time-averaged power for each of the summations as a pixel in the resulting beamforming image (see Figure 4.10). White Gaussian noise (WGN) can be added to the data at this stage to simulate more realistic signal-to-noise ratios (SNR) if desired. The beamforming peak maxima is expected to have a Bessel function shape due to the circular symmetry of the array, and by analyzing the size of the beamforming maxima one can confirm that the beamforming reconstruction measurement has similar position resolution as expected from Locust simulations. Additionally, signal detection rates can be estimated from the data by comparing the magnitude of the beamforming signal peak in the frequency spectra to simulation.

## FSCD Beamforming Simulations

Using Locust simulations of the FSCD one can perform beamforming reconstruction studies using the simulated CRES signal data. As we mentioned in the previous section, the beamforming procedure beings by specifying a set of beamforming positions and corresponding beamforming shifts. The beamforming positions form a grid that covers the region of interest in the field of view of the antenna array. There are effectively an infinite number of ways to specify the grid positions, however, uniform square grids are the most commonly used due to their simplicity. In the FSCD experiment the number and pattern of the grid positions would be optimized to cover the most important regions of the trap volume to maximize detection efficiency while minimizing superfluous calculations.

The beamforming grids used for signal reconstruction with the FSCD consist of a set of points that cover a region of the two-dimensional plane formed by the perimeter of the antenna array. The axial dimension is left out of the beamforming grid because the electrons are assumed to occupy only an average axial position, which corresponds to the center of the magnetic trap. This is because it is impossible to resolve the axial position of the electron as a function of time due to the rapid axial oscillation frequencies of trapped electrons relative to the FSCD time-slice duration.

After beamforming, a summed time-series is obtained for each beamforming position that can be evaluated for the presence of a signal using a detection algorithm. A beamforming image is a visualization method that is equivalent to arranging the beamforming grid points according to their physical locations to form a three-dimensional matrix where the first two dimensions encode the XY-position of the beamforming point and

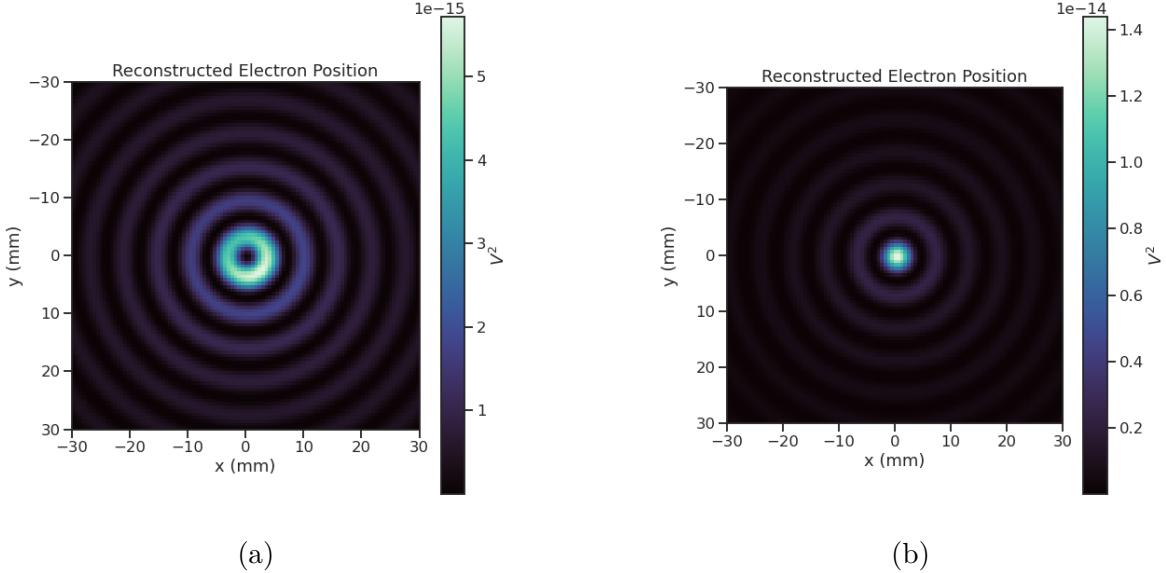


Figure 4.11: Beamforming images visualizing the reconstruction of an electron without (a) and with (b) the cyclotron phase correction. The images were generated using data from Locust simulations. The cyclotron phase refers to a phase offset equal to the relative azimuthal position of an antenna in the array. This phase offset is caused by the circular electron orbit and must be corrected for during reconstruction.

2009 the third dimension contains the summed time-series. The image is formed by taking the  
 2010 time-averaged power (see Figure 4.11). Beamforming images are purely for the purposes  
 2011 of visualization and are not particularly useful for signal detection or reconstruction.

2012 If the beamforming phases consist only of the spatial phase component from Equation  
 2013 4.20, then the resulting beamforming image contains a relatively high-power ring-shaped  
 2014 region that is centered on the position of the electron (see Figure 4.11a). The origin  
 2015 of this shape is an additional phase offset particular to a cyclotron radiation source.  
 2016 Essentially, the circular motion that produces the cyclotron radiation introduces a relative  
 2017 phase offset to the electric fields that is equal to the azimuthal position of the field  
 2018 measurement point. For example, if we have two antennas, one located at an azimuthal  
 2019 position of  $0^\circ$  and another located at an azimuthal position of  $90^\circ$ , then the CRES signals  
 2020 received by these antennas will be out of phase by  $90^\circ$ , which is the difference in their  
 2021 azimuthal positions. This phase offset can be corrected by adding an additional term to  
 2022 the beamforming phase equation that is equal to the azimuthal position of the antenna  
 2023 relative to the electron,

$$\phi_i[n] = \frac{2\pi d_i[n]}{\lambda} + \Delta\varphi_i[n], \quad (4.23)$$

2024 where  $\Delta\varphi_i$  is difference between the azimuthal position of the electron and the  $i$ -th

2025 antenna channel. Using the updated beamforming phases in the summation changes the  
 2026 ring feature into a Bessel function peak whose maximum corresponds to the position of  
 2027 the electron. Including this cyclotron phase correction significantly improves the signal  
 2028 detection and reconstruction capabilities of beamforming by more than doubling the  
 2029 summed signal power and shrinking the beamforming maxima feature size.

2030 The beamforming image examples in Figure 4.11 were produced using an electron  
 2031 located on the central axis of the magnetic trap, which do not experience  $\nabla B$ -drift.  
 2032 However, for electrons produced at non-zero radial position the beamforming phases  
 2033 must be made time-dependent in order to track the position of the electron's guiding  
 2034 center over time. Without this correction the  $\nabla B$ -drift causes the electron to move  
 2035 between beamforming positions, which effectively spreads the cyclotron radiation power  
 over a wider area in the beamforming image (see Figure 4.12). This effect significantly

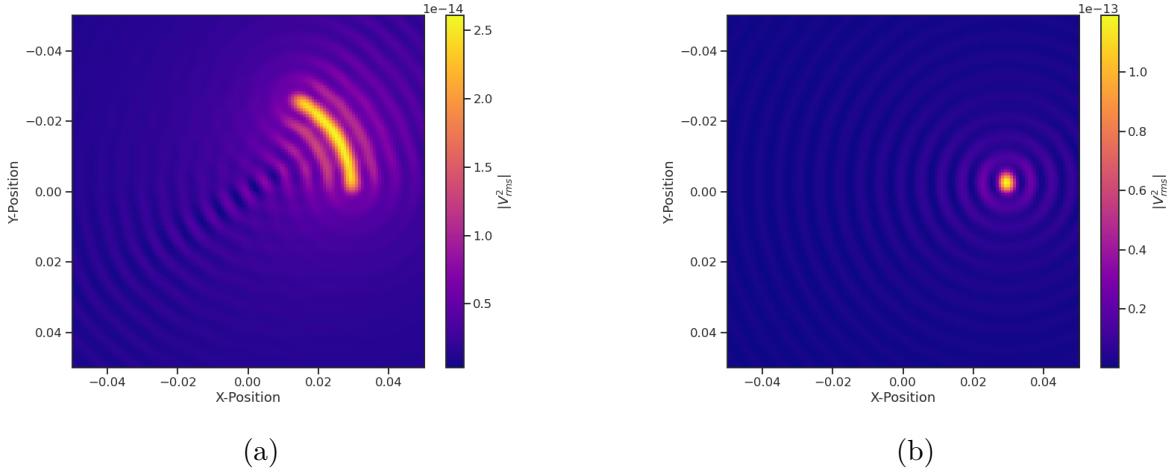


Figure 4.12: Beamforming images visualizing the reconstruction of an electron located off the central axis of the FSCD trap. In (a) we performing beamforming without the  $\nabla B$ -drift correction, and in (b) we include the  $\nabla B$ -drift correction.

2036 reduces the power of the beamforming maxima and increases the size of the beamforming  
 2037 features, simultaneously harming detection efficiency and position reconstruction.  
 2038

2039 The  $\nabla B$ -drift correction simply adds a circular time-dependence to the beamforming  
 2040 positions as a function of time,

$$r[n] = r_0 \quad (4.24)$$

$$\varphi[n] = \varphi_0 + \omega_{\nabla B} t[n], \quad (4.25)$$

2041 where  $\omega_{\nabla B}$  is the drift frequency and  $t[n]$  is the time vector. In the ideal case the  $\nabla B$ -drift

2042 frequencies from Figure 4.2 for the correct pitch angle and radial position would be used,  
2043 however, it is not possible to know the electron’s pitch angle a priori. In principle, one  
2044 could perform multiple beamforming summations for a given beamforming position using  
2045 different drift frequencies and choose the one that maximizes the summed power, but  
2046 this approach leads to a huge computational burden that would be impractical for a  
2047 real FSCD experiment. A compromise is to use an average value of  $\omega_{\nabla B}$  obtained by  
2048 averaging over the drift frequencies for electrons of different pitch angle at a particular  
2049 radius. This approach keeps the computational cost of time-dependent beamforming to a  
2050 minimum while still providing a significant increase in the detection efficiency of digital  
2051 beamforming.

## 2052 **Signal Detection with Beamforming and a Power Threshold**

2053 Up to this point we have neglected any specific discussion of how digital beamforming is  
2054 used for signal detection and reconstruction. This is because, strictly speaking, digital  
2055 beamforming consists only of the phased summation of the array signals and cannot  
2056 be used alone for signal detection. The example beamforming images shown in Figure  
2057 4.11 and Figure 4.12 were produced using simulated data that contained no noise, which  
2058 significantly degrades the utility of analyzing the beamforming images for signal detection  
2059 and reconstruction.

2060 Digital beamforming as a detection algorithm is understood to mean digital beam-  
2061 forming plus a detection threshold placed on the amplitude of the frequency spectrum  
2062 obtained by applying a fast Fourier transform (FFT) to the summed time-series (see  
2063 Figure 4.13). This approach is most similar to the time-frequency spectrogram analysis  
2064 employed in previous CRES experiments, however, in principle any signal detection  
2065 algorithm could be used after the beamforming procedure. In Section 4.4 I analyze the  
2066 signal detection performance of the power threshold approach in detail.

2067 From the example frequency spectra in Figure 4.13 it is clear that without a re-  
2068 construction technique that coherently combines the signals from the full antenna our  
2069 ability to detect CRES signals will be drastically reduced. Because the CRES signals are  
2070 in-phase at the correct beamforming position the summed power increases as a function  
2071 of  $N^2$  compared to a single antenna channel, where  $N$  is the number of antennas. It  
2072 is true that the noise power is also increased by beamforming, but, because the noise  
2073 is incoherent, its power only increases linearly. Consequently, the signal-to-noise ratio  
2074 (SNR) of the CRES signal increases linearly with the number of antennas, which greatly  
2075 improves detection efficiency compared to using only the information in a single antenna.

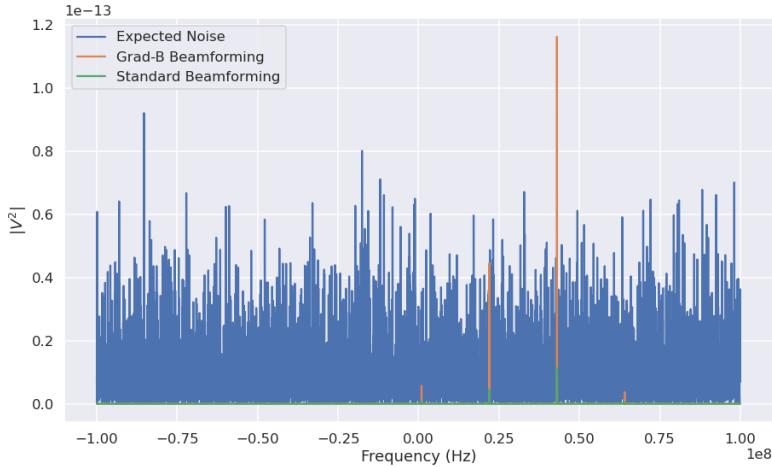


Figure 4.13: A plot of a typical frequency spectrum obtained by applying a Fourier transform to the time-series obtained from beamforming. The frequency spectra are plotted without noise on top of an example of a typical noise spectrum to visualize a realistic signal-to-noise ratio. In the example we see that without beamforming it would not be possible to detect anything since the signal amplitudes would be reduced by a factor of sixty relative to the noise. Additionally, we see that the  $\nabla B$ -drift correction is needed to detect this electron since it comes from a simulation of an electron with a significant off-axis position.

2076     The power threshold detection algorithm searches for high-power frequency bins that  
 2077     should correspond to a frequency component of the CRES signal. In order to prevent  
 2078     random noise fluctuations from being mistaken as CRES signals the power threshold  
 2079     must be set high enough so that it is unlikely that random noise could be responsible. A  
 2080     consequence of this is that many electrons that can be trapped will go undetected because  
 2081     the modulation caused by axial oscillations leads to the cyclotron carrier power to falling  
 2082     below the decision threshold. The time-dependent beamforming used to correct for the  
 2083      $\nabla B$ -drift increases the volume of the magnetic trap where electrons can be detected,  
 2084     but it is ineffective at increasing the range of detectable pitch angles (see Figure 4.14).  
 2085     Fundamentally, this is because the power threshold only uses a fraction of the signal  
 2086     power to detect electrons and ignores the power present in the frequency sidebands. In  
 2087     the subsequent sections I examine two other signal detection algorithms that seek to  
 2088     improve the detection efficiency of the FSCD by utilizing the more of the signal shape to  
 2089     compute the detection test statistics.

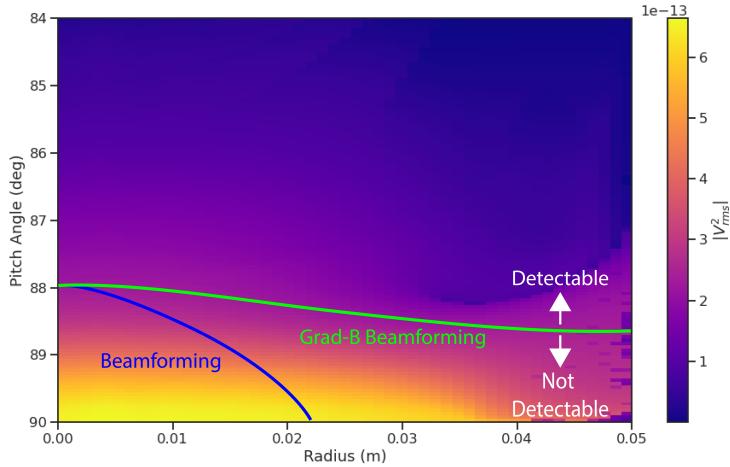


Figure 4.14: A plot of the total signal power received by the FSCD array from trapped electrons with different radial positions and pitch angles generated using Locust simulations. The lines on the plot indicate a 10 dB detection threshold above the mean value of the noise in the frequency spectrum. With static beamforming electrons with radial positions larger than about two centimeters are undetectable due to the change in the electron's position over time causing losses from beamforming phase mismatch. This is corrected by including  $\nabla B$ -drift frequencies in the beamforming phases. Both beamforming techniques fail to detect electrons below  $\approx 88.0^\circ$ , since these signal are composed of several relatively weak sidebands that are comparable to the noise.

### 2090 4.3.2 Matched Filtering

#### 2091 Introduction to Matched Filtering

2092 The problem of CRES signal detection is the problem of detecting a signal buried in  
 2093 WGN, which has been examined at great depth in the signal processing literature [34].  
 2094 For a fully known signal in WGN the optimal detector is the matched filter, which means  
 2095 that it achieves the highest true positive rate for a fixed rate of false positives. The  
 2096 matched filter test statistic is calculated by taking the inner product of the data with  
 2097 the matched filter template

$$\mathcal{T} = \left| \sum_n h^\dagger[n] y[n] \right|, \quad (4.26)$$

2098 where  $h[n]$  is the matched filter template and  $y[n]$  is the data. The matched filter test  
 2099 statistic defines a binary hypothesis test in which the data vector is assumed to be an  
 2100 instance of two possible data classes. By setting a decision threshold on the value of  $\mathcal{T}$ ,  
 2101 one can classify a given data vector as belonging to two distinct hypotheses. Under the  
 2102 first hypothesis the data is composed of pure WGN, and under the second hypothesis the

2103 data is composed of the known signal with additive WGN. The matched filter template  
 2104 is obtained by rescaling the known signal in the following way

$$h[n] = \frac{x[n]}{\sqrt{\tau \sum_n x^\dagger[n]x[n]}}, \quad (4.27)$$

2105 where  $\tau$  is the variance of the WGN and  $x[n]$  is the known signal. Strictly speaking,  
 2106 Equation 4.27 is only true for noise with a diagonal covariance matrix, however, in the  
 2107 context of the FSCD we are justified in assuming this to be true. Defining the matched  
 2108 filter templates in this way guarantees that the expectation value of  $\mathcal{T}$  is equal to one  
 2109 when the data contains only noise, which is the standard matched filter normalization in  
 2110 the signal processing literature.

2111 Although matched filters are canonically formulated in terms of a perfectly known  
 2112 signal, it is still possible to apply the matched filter technique given imperfect information  
 2113 about the signal provided that the signal is deterministic. From our discussion of CRES  
 2114 simulation tools for the FSCD (see Section 4.2) we know that the shape of CRES signals  
 2115 are completely determined by the initial parameters of the electron. The random collisions  
 2116 with background gas molecules which cause the formation of signal tracks are the only  
 2117 stochastic component of the CRES event after the initial beta-decay, therefore, it is  
 2118 possible to develop a matched filter for the detection of CRES signal tracks which are fully  
 2119 determined by the parameters of the electron after the initial beta-decay or subsequent  
 2120 collision events.

2121 The matched filter test statistic for CRES signals is a modified version of Equation  
 2122 4.26

$$\mathcal{T} = \max_{\mathbf{h}, m} |\mathbf{h} * \mathbf{y}| = \max_{\mathbf{h}, m} \left| \sum_k h^\dagger[k]x[m - k] \right|, \quad (4.28)$$

2123 where the matched filter inner product has been replaced with a convolution operation  
 2124 and a maximization over the template and convolution delay ( $m$ ). Replacing the inner  
 2125 product with a convolution accounts for the fact that the start time of the CRES signal is  
 2126 now an unknown parameter, in addition, we now perform a maximization of the matched  
 2127 filter convolution over a number of different templates. Because the shape of the signal is  
 2128 unknown we are forced to guess a number of different signal shapes to create a template  
 2129 bank with which we can identify unknown signals by performing an exhaustive search.

2130 The template bank approach to matched filtering, while quite powerful, can quickly  
 2131 become computationally intractable. This is especially true in the case of the FSCD  
 2132 because of the large amount of raw data produced by the array that must be analyzed.

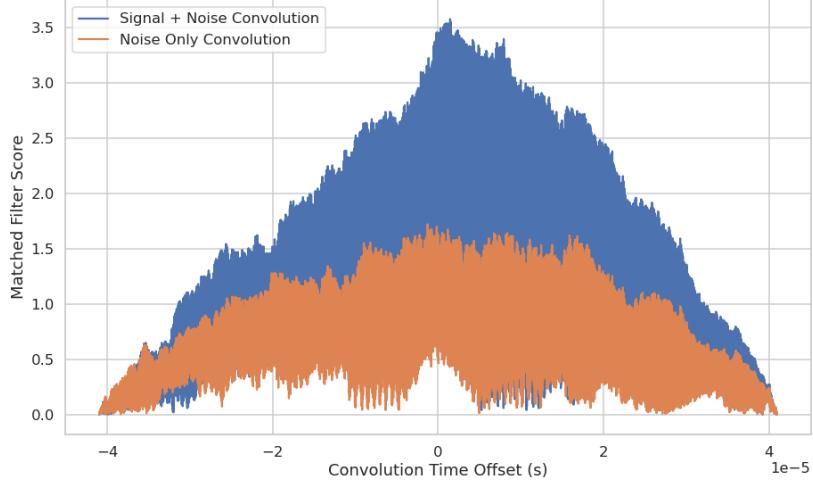


Figure 4.15: Example of a convolution of a CRES signal template with a segment of noisy data. A simulated CRES signal was simulated using Locust and normalized to create a matched filter template. When this template is convolved with noisy data the contains the matching signal the convolution output increases dramatically compared to data with only noise. The decreasing convolution output as the time offset of the convolution increases is caused by zero-padding of the data and template.

2133 Specifically, the time-domain convolution specified by Equation 4.28 is particularly  
 2134 computationally intensive and is a major barrier towards the implementation of a  
 2135 matched filter for signal detection in an experiment like the FSCD. This can be avoided  
 2136 by using the convolution theorem to replace the time-domain convolution with an inner  
 2137 product in the frequency domain.

2138 The convolution theorem states that

$$\mathbf{f} * \mathbf{g} = \mathcal{F}^{-1}(\mathbf{F} \cdot \mathbf{G}) \quad (4.29)$$

2139 where  $\mathbf{f}$  and  $\mathbf{g}$  are discretely sampled time-series,  $\mathbf{F}$  and  $\mathbf{G}$  are the respective discrete  
 2140 Fourier transforms, and  $\mathcal{F}^{-1}$  is the inverse discrete Fourier transform operator. The  
 2141 convolution theorem allows us to perform the matched filter convolution by first com-  
 2142 puting the Fourier transform of the template and data, then performing a point-wise  
 2143 multiplication of the two frequency series, and finally performing the inverse Fourier  
 2144 transform to obtain the convolution output. Because discrete Fourier transforms can be  
 2145 performed extremely efficiently, the convolution theorem is almost always used in lieu of  
 2146 directly computing the convolution.

2147 One thing to note here is that the convolution theorem for discrete sequences shown

here, is technically valid only for circular convolutions, which is not directly specified in Equation 4.28. However, because typical CRES track lengths are much longer than the Fourier analysis window and also that the frequency chirp rates are small compared to the time-slice duration, it is relatively safe to use circular convolutions to evaluate matched filter scores for CRES signals, which allows us to apply the convolution theorem to compute matched filter scores using the frequency representation of the data and matched filter template.

## Matched Filter Analysis of the FSCD

The optimality provided by the matched filter makes it a useful algorithm for analysis of CRES experiment designs for sensitivity analyses, since it indicates the best possible detection efficiency achievable by an experiment configuration. The standard approach to performing these studies involves generating a large number of simulated electron signals that span the kinematic parameter space of electrons in the magnetic trap. In general, electrons have six kinematic parameters along with an additional start time parameter.

In order to limit the number of simulations required to evaluate the detection efficiency the standard approach is to fix the starting axial position, starting azimuthal position, starting direction of the perpendicular component of the electron's momentum, and event start time to reduce the parameter space to starting radial position, starting kinetic energy, and starting pitch angle. The fixed variables are true nuisance parameters that do not affect the detection efficiency estimates for the FSCD design, because they manifest as phases which are marginalized during the calculation of the matched filter score.

Across radial position, kinetic energy, and pitch angle one defines a regular grid of parameters and uses Locust to simulate the corresponding signals (see Figure 4.16). This grid of simulated signals can be used to estimate the likelihood of detecting signals, because the matched filter score specifies the shape of the PDF that defines the detection probability and the size of the template bank influences the likelihood of a good match between a template and a random signal.

The matched filter approach can also be used to estimate the achievable energy resolution of the experiment by using a dense grid of templates generated with parameters close to the unknown signal (see figure 4.17). Because matched filter templates with similar parameters have signal shapes that are also similar, templates with incorrect parameters can have nearly identical matched filter scores as the correct template. Since only one sample of noise is included in a sample of real data, one cannot guarantee that the best matching template corresponds to the ground truth parameters of the signal.

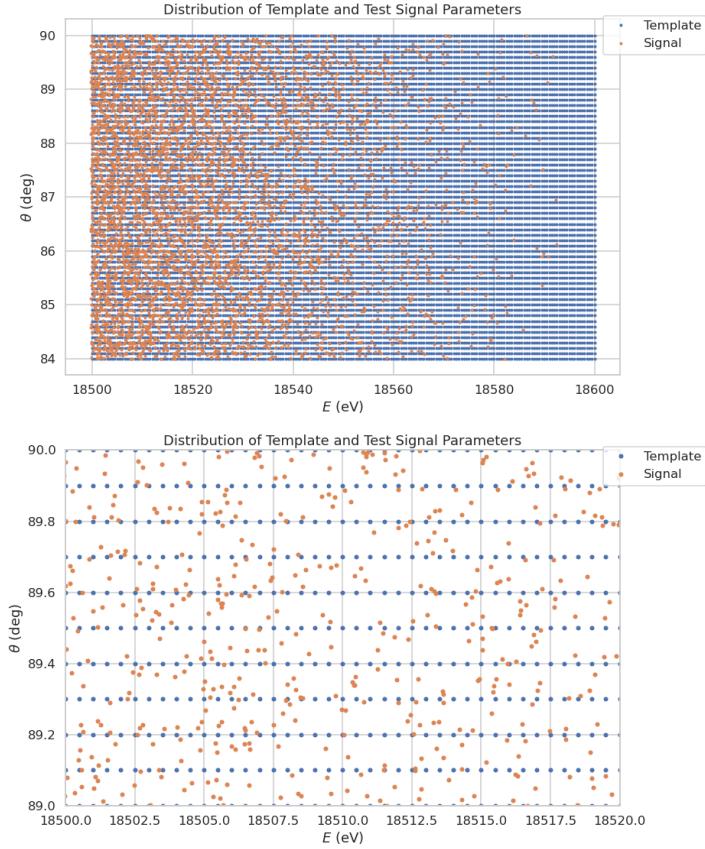


Figure 4.16: An example two-dimensional parameter distribution of a matched filter template bank and random test signals.  $\theta$  refers to the pitch angle of the electron and  $E$  is the kinetic energy. The template bank forms a regular grid of in pitch angle and energy, whereas, the test signals are uniformly distributed in pitch angle and follow the tritium beta-decay kinetic energy distribution. This is why there are fewer test signals at higher energies. The need for high match across the full parameter space prevents one from reducing the density of templates in this low activity region. A zoomed in version of the template bank illustrates the relative density of templates and signals needed for match  $> 90\%$ .

This introduces uncertainty into the signal parameter estimation that manifests as an energy broadening. Dense grids of matched filter templates allows one to quantify this broadening by analyzing the parameter space of templates with matched filter scores close to the ground truth. This approach is analogous to maximum likelihood estimation and is one key component of a complete sensitivity analysis for an antenna array CRES experiment.

A key parameter for describing the performance of a matched filter template bank at signal detection is match, which we define as the average ratio of the highest matched

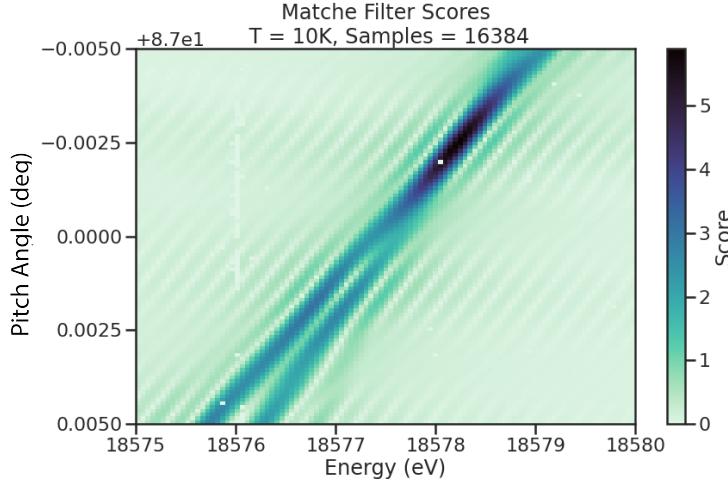


Figure 4.17: The matched filter scores of a dense grid of templates in pitch angle energy space. Dense template grids allow one to estimate the kinetic energy of the electron by identifying the best matching template. The uncertainty on this value is proportional to the space of templates that also match the test signal well. In the worst case matched filter templates can be completely degenerate where templates with different parameters match a signal with equal likelihood.

2190 filter score for a random signal to the matched filter score for a perfectly matching  
2191 template. In equation form this is

$$\text{Match} \equiv \Gamma = \frac{\mathcal{T}_{\text{best}}}{\mathcal{T}_{\text{ideal}}}, \quad (4.30)$$

2192 where  $\mathcal{T}_{\text{best}}$  is the matched filter score of the best fitting template in the bank and  $\mathcal{T}_{\text{ideal}}$  is  
2193 the hypothetical matched filter score one would measure if the signal perfectly matched  
2194 the template. Generally, one desires an average match as close to one as possible, however,  
2195 the average match value is an exponential function of the number of templates in the  
2196 template bank (see Figure 4.18). This behavior is observed for dense matched filter grids  
2197 like the one in Figure 4.17. A dense grid was used to calculate the average value of match  
2198 for different template bank sizes shown in Figure 4.18.

2199 The exponential relationship between match and template bank size is also evident  
2200 for template banks that cover a wide range of parameters, such as the template bank  
2201 visualized in Figure 4.16. Since no prior knowledge of the signal parameters is available,  
2202 one has no choice but to use a template bank that covers a large range of parameters for  
2203 signal detection. Achieving a high average match in this scenario can easily overwhelm  
2204 the available computational resources, so in practice only a limited number of templates

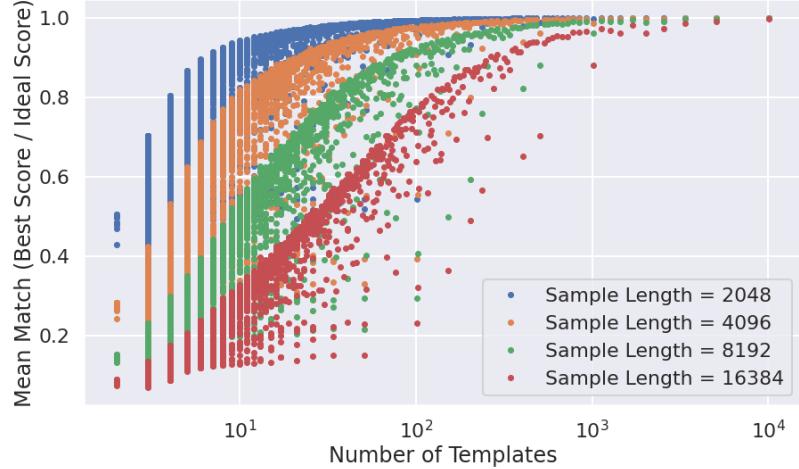


Figure 4.18: The mean match of the dense template grid shown in Figure 4.17 for different numbers of templates. Grids of different sizes were obtained by decimating a dense grid of templates and the average match for each grid was computed using the same set of randomly distributed test signals. Plotting the mean match against the size of the grid allows one to visualize the exponential relationship between match and template bank size. The noise in each curve is caused by sampling effects from the decimation algorithm. In general, longer templates are harder to than shorter templates.

2205 could be used at the detection stage. Therefore, accurately modeling the effects of match  
2206 is key to correct sensitivity calculations.

2207 The effect of match on the detection efficiency of the matched filter template bank can  
2208 be summarized using the ROC curve (see Figure 4.19). A single ROC curve is obtained  
2209 by averaging over the PDFs that describe the detection probabilities of each individual  
2210 template. The matched filter score for a template follows a Rician distribution with a  
2211 mean value equal to the matched filter score multiplied by the match ratio between the  
2212 template and signal. Therefore, the distribution that describes the average matched filter  
2213 score when there is a signal in the data is obtained by averaging over the distributions  
2214 for every template, whose expectation values are multiplied by the average match ratio.

2215 The distribution of the matched filter score when there is no signal in the data follows  
2216 a Rayleigh distribution. Therefore, a trials penalty, which is the statistical penalty one  
2217 pays for randomly checking many templates in order to avoid a random match between  
2218 noise and a template, is included by computing the joint distribution of  $N_{\text{template}}$  Rayleigh  
2219 distributions, where  $N_{\text{template}}$  is the size of the template bank. For more information on  
2220 the calculation of matched filter template bank ROC curves please refer to Section 4.4.

2221 An alternative way to visualize the detection performance for each algorithm is to

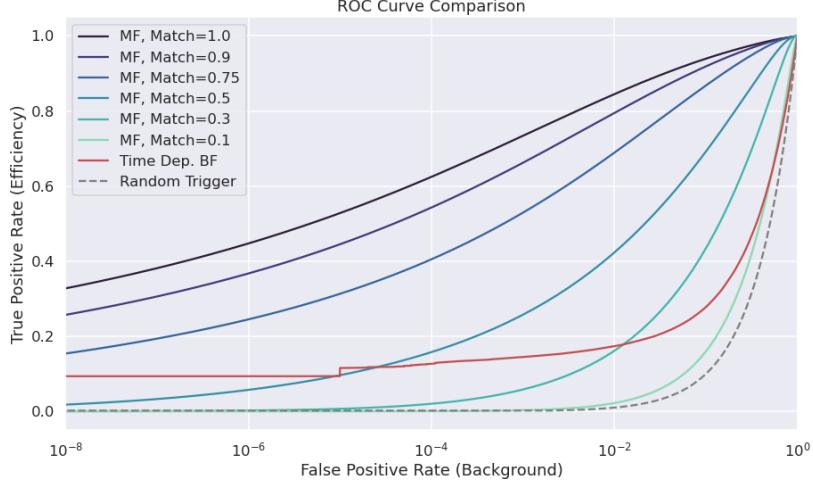


Figure 4.19: Matched filter template bank ROC curves as a function of mean match. One can see that for low match a matched filter is on average worse than the more straight forward beamforming detection approach.

2222 specify a minimum acceptable false positive rate at the trigger level. This is equivalent  
 2223 to specifying a minimum threshold on the value of the matched filter score or the size of  
 2224 a frequency peak for a beamforming power threshold trigger. One can then draw regions  
 2225 of detectable signals as a function of the electron's pitch angle and radial position (see  
 2226 Figure 4.20). A kinetic energy shift is equivalent to an overall frequency shift of the  
 2227 signal and should have no effect on the detection probability assuming sufficient density  
 2228 of matched filter templates in the energy dimension. A electron is declared "detectable"  
 2229 for the regions in Figure 4.20 if the signal has at least 50% probability of falling above the  
 2230 decision threshold of the respective classifier. One can see that the parameter space of  
 2231 detectable signals is greatly expanded beyond the beamforming power threshold trigger  
 2232 with a matched filter (MF) or deep neural network (DNN) (see Section 4.3.3). Plots such  
 2233 as Figure 4.20 are useful for visualization, but, since the handling of detection likelihood  
 2234 is not sufficiently rigorous, the detection probability boundaries are not well-suited to  
 2235 sensitivity estimates.

### 2236 Optimized Matched Filtering Implementation for the FSCD

2237 The biggest practical obstacle to the implementation of a matched filter template bank  
 2238 detection approach is oftentimes the computational cost associated with exhaustively  
 2239 calculating the matched filter scores of the template bank, and the FSCD is no exception  
 2240 in this regard. At a basic level computing a matched filter score requires the convolution

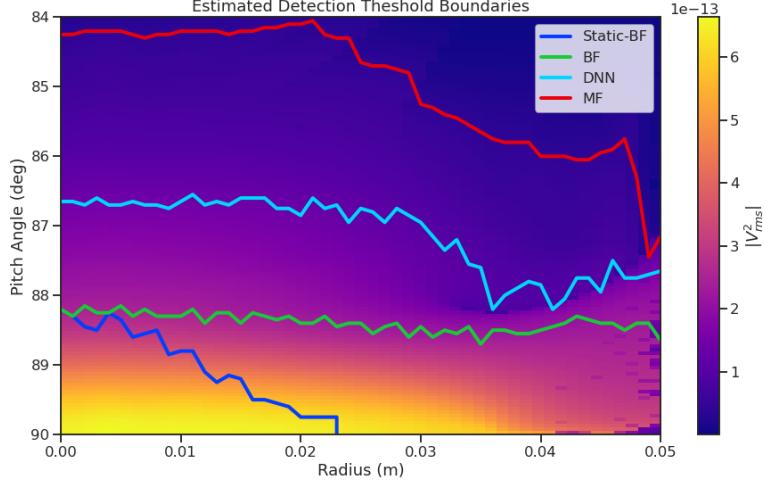


Figure 4.20: Boundaries of detectable electrons in pitch angle kinetic energy space for a series of different signal detection algorithms. A detectable signal is defined as a signal that is above a consistent decision with at least 50% probability. This non-rigorous treatment of detection probability is primarily useful for the visualization the relative increases in detection performance provided by the different algorithms. The static beamforming (Static-BF) algorithm is the digital beamforming algorithm introduced above without the  $\nabla B$ -drift correction. The DNN algorithm refers to a convolutional neural network classifier trained to detect CRES signals (see Section 4.3.3).

of two vectors, which can be performed very efficiently by computers if the convolution theorem and fast Fourier transforms (FFT) are utilized. Furthermore, one can consider applying digital beamforming as a pre-processing step to reduce the dimensionality of the data before the matched filter is applied. In order to understand the relative gain in computational efficiency offered by these optimizations we analyze the total number of floating-point operations (FLOP) of several matched filter implementations in big  $O$  notation that utilize different combinations of optimizations.

A direct implementation of a matched filter as specified by Equation 4.28 involves the convolution of  $N_{\text{ch}}$  signals of length  $N_s$  with template signals of length  $N_t$ . As a uniform metric we shall compare the FLOP of the various matched filter implementations on a per-template basis, since each implementation scales linearly with the number of templates. The direct convolution approach to matched filtering costs

$$O(N_{\text{ch}}) \times O(N_s \times N_t) \quad (4.31)$$

FLOP per-template, whose cost is dominated by the  $O(M \times N)$  convolution operation. The computational cost of the direct matched filter approach can be significantly

2255 reduced by exploiting the convolution theorem and FFT algorithms. If we restrict  
 2256 ourselves to signals and templates that contain equal numbers of samples then the  
 2257 convolution can be calculated by Fourier transforming both vectors, performing the  
 2258 point-wise multiplication, and then performing the inverse Fourier transform to obtain  
 2259 the convolution result. The FFT algorithm is able to compute the Fourier transform  
 2260 utilizing only  $O(N \log N)$  operations compared to  $O(N^2)$  for a naive Fourier transform  
 2261 implementation. This optimization results in a computational cost per-template of

$$O(N_{\text{ch}}) \times O(N_s \log N_s) \quad (4.32)$$

2262 A typical signal vector in the FSCD contains  $O(10^4)$  samples in which case the FFT  
 2263 reduces the computational cost of the matched filter by a factor of  $O(10^3)$ . This large  
 2264 reduction in computational cost implies that a direct implementation of a matched filter  
 2265 is completely infeasible in the FSCD due to resource constraints.

2266 Rather than relying solely on the matched filter it is tempting to consider using  
 2267 digital beamforming as an initial step in the signal reconstruction for the purposes of  
 2268 data reduction. The primary motivation is to reduce the dimensionality of the data by  
 2269 a factor of  $N_{\text{ch}}$  by combining the array outputs coherently into a single channel. One  
 2270 can view the beamforming operation as a partial matched filter, in the sense that the  
 2271 matched filter convolution contains the beamforming phased summation along with a  
 2272 prediction of the signal shape. By separating beamforming from the signal shape one  
 2273 hopes to reduce the overall computational cost by effectively shrinking the number of  
 2274 templates and reducing the number of operations required to check each one.

2275 The nature of this optimization requires that we account for the number of templates  
 2276 used for pure matched filtering versus the hybrid approach. To first order, the total  
 2277 number of templates at the trigger stage is a product of the number of guesses for each  
 2278 of the electron's parameters

$$N_T = N_E \times N_\theta \times N_r \times N_\varphi, \quad (4.33)$$

2279 where  $N_E$  is the number of kinetic energies,  $N_\theta$  is the number of pitch angles,  $N_r$  is the  
 2280 number of starting radial positions, and  $N_\varphi$  is the number of starting azimuthal positions.  
 2281 The starting axial position and cyclotron motion phase are not necessary to include in  
 2282 the template bank since these parameters manifest themselves as the starting phase of  
 2283 the signal, which is effectively marginalized when using a FFT to compute the matched  
 2284 filter convolution. Therefore, the total number of operations required by a matched filter

2285 to detect a signal in a segment of array data is on the order of

$$O(N_T) \times O(N_{ch}) \times O(N_s \log N_s) \quad (4.34)$$

2286 With the hybrid approach we attempt to remove the spatial parameters from the  
 2287 template bank by using beamforming to combine the array signals into a single channel.  
 2288 Beamforming explicitly assumes a starting position, which allows us to only use matched  
 2289 filter templates that span the two-dimensional space of kinetic energy and pitch angle.  
 2290 The total computational cost of the hybrid method is directly proportional to the number  
 2291 of beamforming positions. For the time-dependent beamforming defined in Section 4.3.1,  
 2292 the number of beamforming positions is given by

$$N_{BF} = N_r \times N_\varphi \times N_{\omega_{\nabla B}}, \quad (4.35)$$

2293 where  $N_r$  and  $N_\varphi$  are the same spatial parameters encountered in the pure matched  
 2294 filter template bank and  $N_{\omega_{\nabla B}}$  is the number of  $\nabla B$ -drift frequency assumptions. If a  
 2295 unique drift frequency is used for each pitch angle then the hybrid approach is effectively  
 2296 equivalent to a pure matched filter in the number of operations. The key efficiency gain  
 2297 of the hybrid approach is to exploit the relatively small differences in  $\omega_{\nabla B}$  for electrons  
 2298 of different pitch angles by using only a small number of average drift frequencies.

2299 The total number of operations for the hybrid approach can be expressed as a sum of  
 2300 the operations required by the beamforming and matched filtering steps,

$$O(N_{BF}) \times O(N_{ch} N_s) + O(N_{BF}) \times O(N_E N_\theta) \times O(N_s \log N_s). \quad (4.36)$$

2301 The first product in the sum is the number of operations required by beamforming,  
 2302 which is simply the number of beamforming points times the computational cost of the  
 2303 beamforming matrix multiplication, and the second product is the computational cost  
 2304 of matched filtering the summed signal generated by each beamforming position. To  
 2305 compare this to pure matched filtering we take the ratio of Equations 4.34 and 4.36 to  
 2306 obtain

$$\Gamma_{BFMF} = \frac{O(N_{\omega_{\nabla B}})}{O(N_E N_\theta) \times O(\log N_s)} + \frac{O(N_{\omega_{\nabla B}})}{O(N_{ch})}. \quad (4.37)$$

2307 This expression can be simplified by observing that  $O(N_E N_\theta) \times O(\log N_s) \gg O(N_{ch})$ ,

2308 which means that the ratio of computational cost for the two methods can be reduced to

$$\Gamma_{\text{BFMF}} \approx \frac{O(N_{\omega_{\nabla B}})}{O(N_{\text{ch}})}. \quad (4.38)$$

2309 If we limit ourselves to a number of estimated drift frequencies of  $O(1)$  then we see that  
2310 the estimated computational cost reduction of the hybrid approach is of  $O(N_{\text{ch}})$ . This is  
2311 quite a large reduction considering that the FSCD antenna array contains sixty antennas  
2312 in the baseline design.

2313 The main drawback of the hybrid approach is that the limited number of allowed  
2314 drift frequency guesses can lead to detection efficiency loss due to phase mismatch. The  
2315 degree of phase error from an incorrect drift frequency is proportional to the length of  
2316 the array data vector used by the signal detection algorithm. For signals with lengths  
2317 equal to the baseline FSCD Fourier analysis window of 8192 samples, typical phase errors  
2318 from using an average versus the exact  $\nabla B$ -drift frequency are on the order of a few  
2319 percent in terms of the signal energy. This has a relatively small impact on the overall  
2320 detection efficiency, however, future experiments with antenna array CRES will want to  
2321 balance optimizations such as these during the design phase to keep experiment costs to  
2322 a minimum while still achieving scientific goals.

### 2323 Kinetic Energy and Pitch Angle Degeneracy

2324 More accurate modeling of a matched filter requires that we consider the effects of  
2325 mismatched signals and template, since this more accurately reflects the real-world usage  
2326 of a matched filter where many incorrect templates are convolved with the data until the  
2327 matching template is found. One way to study this is to use the grid of simulated signals  
2328 to compute the matched filter scores between mismatched signals and templates and  
2329 evaluate the matched filter scores under this scenario. What one finds when performing  
2330 this analysis is that templates for kinetic energies and pitch angles that do not match  
2331 the underlying signal can have matched filter scores that are indistinguishable from the  
2332 matched filter score of the correct template (see Figure 4.21 and Figure 4.21).

2333 This degeneracy in matched filter score is the result of correlations between the kinetic  
2334 energy of the electron and the pitch angle caused by changes in the average magnetic field  
2335 experienced by an electron for different pitch angles. While in principle helpful for the  
2336 purposes of signal detection these correlations are unacceptable since they greatly reduce  
2337 the energy resolution of the experiment by causing electrons with specific kinetic energy  
2338 to templates across a wide range of energies. It is important to emphasize that this

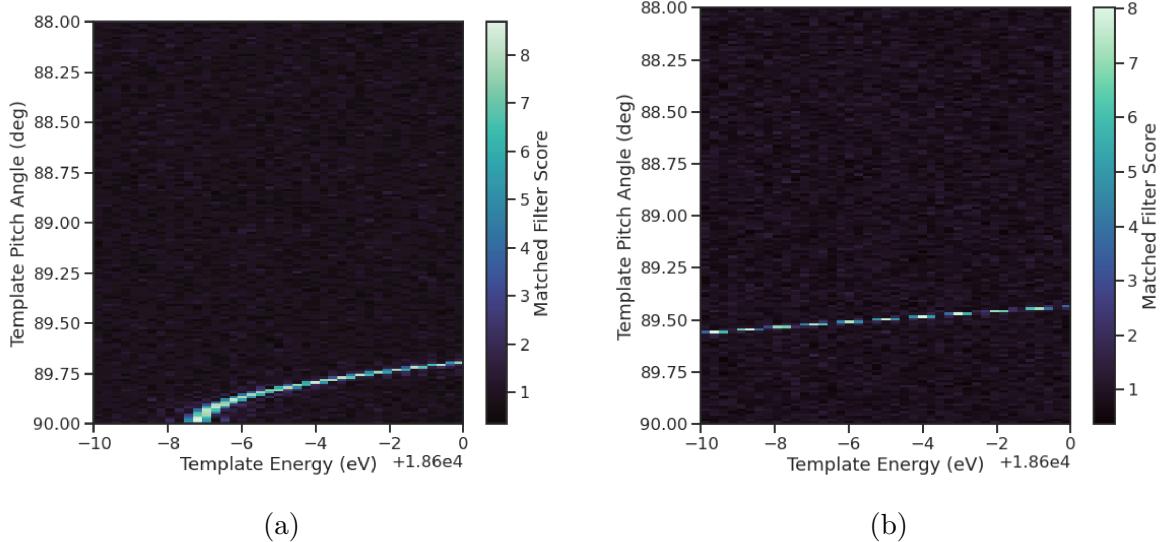


Figure 4.21: Two example illustrations of the correlation between kinetic energy and pitch angle imparted by the shape of the FSCD magnetic trap. The correlations manifest themselves as degeneracies in the matched filter score where multiple matched filter templates have the same matched filter for a particular signal. These degeneracies are a sign that the magnetic trap must be redesigned in order to break the correlation between pitch angle and kinetic energy.

2339 degeneracy cannot be fixed by implementing a different signal reconstruction algorithm.  
2340 As revealed by the matched filter scores the shapes of the signals for different parameters  
2341 are identical. Resolving this degeneracy between pitch angle and energy requires the  
2342 design of a new magnetic trap with steeper walls so that the average magnetic field  
2343 experienced by an electron is less dependent on pitch angle.

### 2344 4.3.3 Machine Learning

2345 Machine learning is a vast and rapidly developing field of research [37]. In this Section  
2346 we shall attempt to provided a brief introduction to some of the concepts and techniques  
2347 of machine learning that were applied to CRES signal detection rather than attempt a  
2348 comprehensive overview.

#### 2349 Introduction to Machine Learning

2350 Digitization of the FSCD antenna array generates large amounts of data that must be  
2351 rapidly processed to enable real-time signal detection and reconstruction. While digital  
2352 beamforming combined with a power threshold is relatively computationally inexpensive,

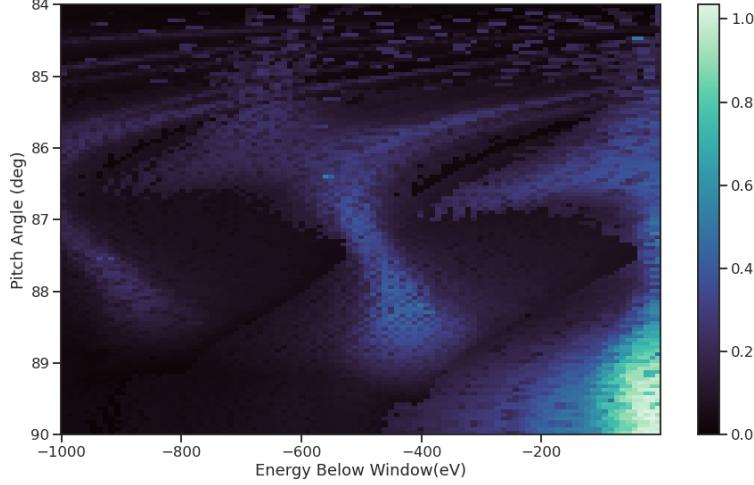


Figure 4.22: A visualization of the correlation between energy and pitch angle in the FSCD magnetic trap. The image is formed by computing the match of the best template from a grid consisting of pitch angles from 84 to 90 degrees in steps of 0.05 degrees, kinetic energies from 17574 to 18574 eV, located at 2 cm from the central axis, and simulated for a length of three FSCD time-slices. The signals used to compute the best matching template consisted of a grid from 84 to 90 degrees in steps of 0.05 degrees, kinetic energies from 18550 to 18575 eV in steps of 0.25 eV, located 2 cm from the central axis, and simulated for three FSCD time-slices. The colored regions of the plot show how well signals with lower energy can match those of higher energy for the FSCD magnetic trap, which is proportional to the achievable energy resolution of the FSCD design.

it is relatively ineffective at detecting CRES signal with small pitch angles, since it relies on a visible frequency peak above the noise. On the other hand, a matched filter is able to detect signals with a significantly larger range of parameters, however, the exhaustive search of matched filter templates can be computationally expensive. Machine learning based triggering algorithms have been used successfully in many different high-energy physics experiments [38] and recent developments have shown success in the detection of gravitational wave signals using machine learning techniques [39, 40] in place of the more traditional matched filtering method. This motivates the exploration of machine learning as a potential CRES signal detection algorithm.

There are several different approaches to machine learning, but the one most important to our discussion here is the supervised learning approach. In supervised machine learning one uses a differentiable model or function that is designed to map the input data to the appropriate label [37]. The data is represented as a multidimensional matrix of floating point values such as an image or a time-series, and the label is generally a class name such as signal or noise for classification problems or a continuous value like kinetic energy

2368 in the case of regression problems.

2369 In supervised learning the model is trained to map from the data to the correct label  
2370 by evaluating the output of the model using a training dataset consisting of a set of  
2371 paired data and labels. To evaluate the difference between the model output and the  
2372 correct label a loss function is used to quantify the error between the model prediction  
2373 and the ground truth. For example, a common loss function in regression problems is the  
2374 squared error loss function, which quantifies error using the squared difference between  
2375 the model output and label.

2376 Using the outputs of the loss function the next step in supervised learning is to  
2377 compute the gradient of error with respect to the model parameters in a process called  
2378 backpropagation. Using the model parameter gradients the last step in the supervised  
2379 learning process is to perform an update of the parameter values in order to minimize  
2380 the error in the model predictions across the whole dataset. This loop is performed many  
2381 times while randomly shuffling the dataset until the error converges to a minimum value  
2382 at which point the training procedure has finished. It is standard practice to monitor  
2383 the training procedure by evaluating the performance of the model using a separate  
2384 validation dataset that matches the statistical distribution of the training data and to  
2385 check the performance of the model after training using yet another dataset called the  
2386 test dataset. These practices help to guard against overtraining which is a concern for  
2387 models with many parameters.

## 2388 Convolutional Neural Networks

2389 A popular class of machine learning models are neural networks. A neural network is  
2390 essentially a function composed of a series of linear operations called layers which take a  
2391 piece of data typically represented as a matrix, multiplies the elements of the data by a  
2392 weight, and then sums these products to produce an output matrix. Neural networks  
2393 composed of purely linear operations are unable to model complex non-linear behavior,  
2394 therefore, non-linear activation functions are applied to the outputs of each of the layers  
2395 to increase the ability of the neural network to model complex relationships between the  
2396 data.

2397 Neural networks are typically composed of at least three layers, but with the present  
2398 capabilities of computer hardware they more often contain many more than this. The  
2399 first layer in a neural network is called the input layer, because it takes the data objects  
2400 as input, and the last layer in a neural network is known as the output layer. The  
2401 output layer is trained by machine learning to map the data to a desired output using

2402 the supervised learning procedure described in Section 4.3.3. In between the input and  
2403 the output layer are typically several hidden layers that receive inputs from and transmit  
2404 outputs to other layers in the neural network model. The term deep neural network  
2405 (DNN) refers to those neural networks that have at least one hidden layer, which have  
2406 proven to be extremely powerful tools for pattern recognition and function approximation.

2407 An important type of DNN are convolutional neural networks (CNN) that typically  
2408 contain several layers which perform a convolution of the input with a set of filters. These  
2409 convolution operations are typically accompanied by layers that attempt to down-sample  
2410 the data along with the standard neural network activation functions. A standard CNN  
2411 is composed of several convolutional layers at the beginning of the network and ends  
2412 with a series of fully-connected neural network layers at the output. Intuitively, one  
2413 can imagine that the convolutional layers are extracting features from the data that  
2414 fully-connected layers use to perform the classification or regression task.

2415 **Deep Filtering for Signal Detection in the FSCD**

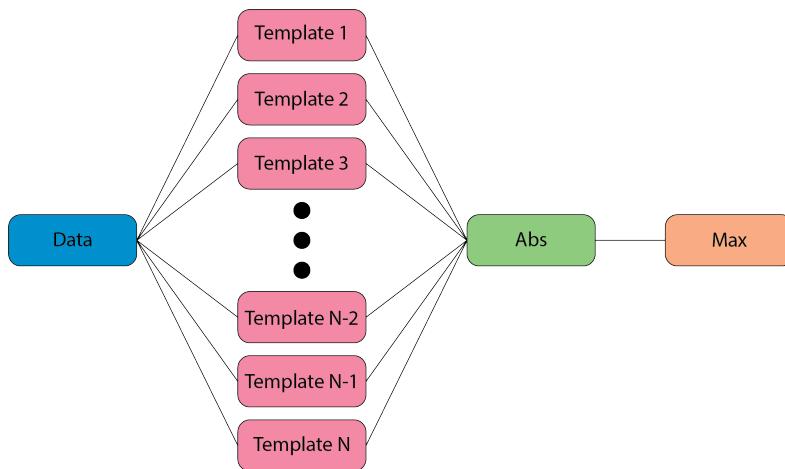


Figure 4.23: A representation of a matched filter template bank as a convolutional neural network. The network has a single layer composed of the templates, which act as convolutional filters. The activation of the neural network is an absolute value followed by a max operator.

2416 CNNs have been extremely influential in the field of computer vision, particularly tasks  
2417 such as image segmentation and classification, but have also been applied in numerous  
2418 experimental physics contexts. Given the particular challenge posed by signal detection  
2419 and reconstruction in the FSCD we are interested in exploring the potential of machine  
2420 learning as an effective algorithm for real-time signal detection, since this application

2421 requires both high efficiency and fast evaluation.

2422 In the machine learning paradigm signal detection is equivalent to a binary classifi-  
2423 cation problem between the signal and noise data classes, and my investigation focuses  
2424 specifically on the application of CNNs to signal detection in the FSCD, which is moti-  
2425 vated by relatively recent demonstrations of CNNs achieving classification accuracies for  
2426 gravitational wave time-series signals comparable to a matched filter template bank. In  
2427 this framework it is possible to interpret the matched filter as a type of CNN composed  
2428 of a single convolutional layer with the templates making up the layer filters (see Figure  
2429 4.23). Since this neural network has no hidden layers, it is not a DNN like we have  
2430 been discussing so far, but we can attempt to construct a proper CNN that attempts to  
2431 reproduce the classification performance of the matched filter network.

2432 The name deep filtering refers to this scheme of replacing a matched filter template  
2433 bank with a DNN. The reason why one might want to do this is that it may be possible to  
2434 exploit redundancies and correlations between templates that may allow one to perform  
2435 signal detection with similar accuracy but with fewer computations, which is important  
2436 for real-time detection scenarios like the FSCD experiment. In Section 4.4 we perform a  
2437 detailed comparison of the signal detection performance of a CNN to beamforming and a  
2438 matched filter template bank.

2439 Deep filtering is conceptually a simple technique. Similar to a matched filter template  
2440 bank a large number of simulated CRES signals are generated and used to train a model  
2441 to distinguish between signal and noise data (see Figure 4.24). In order to reduce the  
2442 dimensionality of the input FSCD data a digital beamforming summation is applied  
2443 to the raw time-series data generated by Locust to compress the 60-channel data to a  
2444 single time-series. CRES signal have a sparse frequency representation and experiments  
2445 training CNN's on time-series and frequency series data found that models trained on  
2446 frequency spectrum data performed significantly better, therefore, an FFT is applied to  
2447 the summed time-series before being normalized and fed to the classification model.

2448 The data used to train the model consists of an equal proportion of signal and noise  
2449 frequency spectra. Unique samples of WGN are generated and added to the signals during  
2450 training time to avoid have to pre-generate and store large samples of noise data. The  
2451 binary cross-entropy loss function combined with the ADAM optimizer proved effective  
2452 at training the models to classify CRES data. A simple hyperparameter optimization  
2453 was performed by manually tuning model, loss function, and optimizer parameters. The  
2454 model and training loops was implemented in python using the PyTorch deep learning  
2455 framework. Standard machine learning best practices were followed when training the

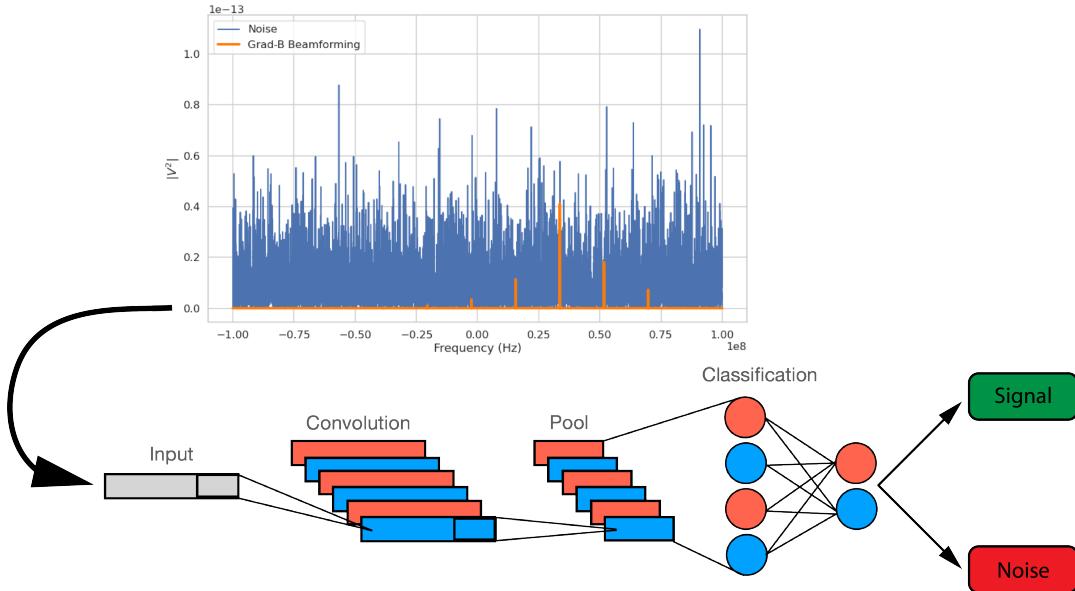


Figure 4.24: A graphical depiction of CRES signal detection using a CNN. A noisy segment of data is converted to a frequency series using digital beamforming and a FFT. The complex-valued frequency series is input into a trained CNN model that classifies the data as signal or noise using a decision threshold on the CNN output.

models, such as overtraining monitoring using a validation dataset. Models were trained until the training loss and accuracy converged and then evaluated using a separate test data set.

The classification results of the test dataset are used to quantify the relationship between the true positive rate and the false positive rate for the model. The true positive rate is analogous to detection efficiency and the false positive rate is a potential source of background in the detector. One can limit the rate of false positives using a sufficiently high threshold on the model output at the cost of a lower detection efficiency (see Figure 4.25 and Figure 4.26). As expected, the performance of the model at signal classification is negatively effected the noise power, which is quantified by the noise temperature.

## 4.4 Analysis of Signal Detection Algorithms for the Antenna Array Demonstrator

This section contains an early version of the manuscript for the triggering paper prepared for publication in JINST. In it I present a relatively detailed analysis of the signal detection performance of the three signal detection approaches discussed so far using a

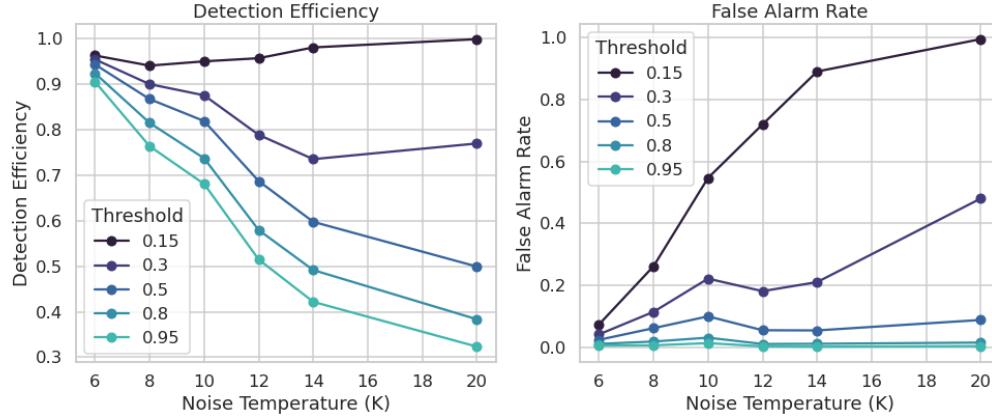


Figure 4.25: The detection efficiency and false alarm rate (false positive rate) as a function of the decision threshold for different values of the noise temperature. The model is trained to output a value close to one for data that contains a signal and outputs a value near zero when the data contains only noise. One sees that a lower decision threshold will have a high detection efficiency at the cost of a high rate of false alarms.

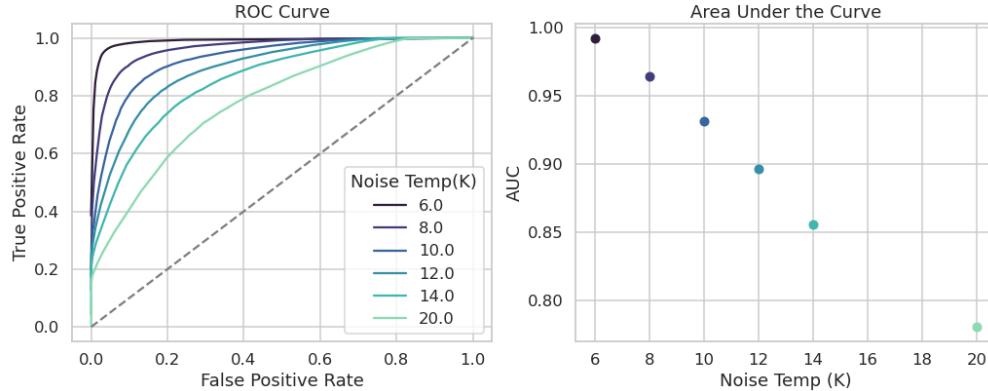


Figure 4.26: ROC curves for a CNN model classifying CRES signals. One can see that the area under the curve, which is a figure of merit that describes the performance of the classifier, is roughly linearly dependent with the noise temperature.

2471 population of simulated CRES signals generated with Locust. The focus of the paper is  
 2472 on the performance of the signal detection algorithms for pitch angles below  $88.5^\circ$  where  
 2473 the beamforming power threshold begins to fail.

#### 2474 4.4.1 Introduction

2475 Cyclotron Radiation Emission Spectroscopy (CRES) is a technique for measuring the  
 2476 kinetic energies of charged particles by observing the frequency of the cyclotron radiation

2477 that is emitted as they travel through a magnetic field [3]. The Project 8 Collaboration  
2478 is developing the CRES technique as a next-generation approach to tritium beta-decay  
2479 endpoint spectroscopy for neutrino mass measurement. Recently, Project 8 has used  
2480 CRES to perform the first ever tritium beta-decay energy spectrum and neutrino mass  
2481 measurement [5, 6].

2482 Previous CRES measurements have utilized relatively small volumes of gas that are  
2483 directly integrated with a waveguide transmission line, which transmits the cyclotron  
2484 radiation emitted by the trapped electrons to a cryogenic amplifier. While this technology  
2485 has had demonstrable success, it is not a feasible option for scaling up to significantly  
2486 larger measurement volumes. In particular, the goal of the Project 8 Collaboration  
2487 is to use CRES combined with atomic tritium to measure the neutrino mass with a  
2488 40 meV sensitivity. Achieving this sensitivity goal will require a multi-cubic-meter scale  
2489 measurement volume in order to obtain the required event statistics in the tritium  
2490 beta-spectrum endpoint region; hence, there is a need for new techniques to enable large  
2491 volume CRES measurements for future experiments.

2492 One approach is to surround a large volume with an array of antennas that together  
2493 collect the cyclotron radiation emitted by trapped electrons [4, 41]. A promising array de-  
2494 sign is an inward-facing uniform cylindrical array that surrounds the tritium containment  
2495 volume. Increasing the size of the antenna array, by adding additional rings of antennas  
2496 along vertical axis, allows one to grow the experimental volume until a sufficient amount  
2497 of tritium gas can be observed by the array. A challenging aspect of this approach is  
2498 that the total radiated power emitted by an electron near the tritium spectrum endpoint  
2499 is on the order of 1 fW or less, which is then distributed between all the antennas in  
2500 the array. Consequently, detecting the presence of a CRES signal and determining the  
2501 electron’s kinetic energy requires reconstructing the entire antenna array output over the  
2502 course of the CRES event, posing a significant data acquisition and signal reconstruction  
2503 challenge.

2504 Project 8 has developed a triggering system to enable real-time identification of CRES  
2505 events using an antenna array [42]. Previous measurements with the CRES technique  
2506 have utilized a threshold on the frequency spectrum formed from a segment of CRES  
2507 time-series data. This algorithm relies on the detection of a frequency peak above the  
2508 thermal noise background, which limits the kinematic parameter space of detectable  
2509 electrons. Due to the limitations of this power threshold, Project 8 has been investigating  
2510 alternative signal identification approaches, including both matched filtering and machine  
2511 learning based classifiers, to improve the detection efficiency of the experiment. In

order to evaluate the relative gains in detection efficiency that come from utilizing these alternative algorithms, we develop analytical models for the power threshold and matched filter signal classifier performance applicable to an antenna array based CRES detector. In addition, we implement and test a basic convolutional neural network (CNN) as a first step towards the development of neural-network based classifiers for CRES measurements. These results allow us to compare the estimated detection efficiencies of each of these methods, which we weigh against the associated computational costs for real-time applications.

The outline of this paper is as follows. In Section 4.4.2 we give an overview of a prototypical antenna array CRES experiment, and describe the major steps involved in the proposed approach to real-time signal identification. In Section 4.4.3 we develop models for the power threshold and matched filter algorithms, and introduce the machine learning approach and CNN architecture. In Section 4.4.4 we describe our process for generating simulated CRES signal data and the details of training the CNN. Finally, in Section 4.4.5 we perform a comparison of the signal classification accuracy of the three approaches and discuss the relevant trade-offs in terms of detection efficiency and computational cost.

## 4.4.2 Signal Detection with Antenna Array CRES

### 4.4.2.1 Antenna Array and DAQ System

In order to explore the potential of antenna array CRES for neutrino mass measurement, the Project 8 Collaboration has developed a conceptual design for a prototype antenna array CRES experiment [4, 41], called the Free-space CRES Demonstrator or FSCD, which could be used as a demonstration of the antenna array measurement technique (see Figure 4.27). The FSCD utilizes a single ring of antennas, which is the simplest form of a uniform cylindrical array configuration, to surround a radio-frequency (RF) transparent tritium gas vessel. A prototype version of this antenna array has been built and tested by the Project 8 collaboration to validate simulations of the array radiation pattern and beamforming algorithms [7]. In the FSCD the antenna array is positioned at the center of the magnetic trap formed by a set of electro-magnetic coils that are designed to produce a magnetic trap with flat central region and steep walls both radially and axially.

When a beta-decay electron is trapped its motion consists of three primary components. The component with the highest frequency is the cyclotron orbit whose frequency is

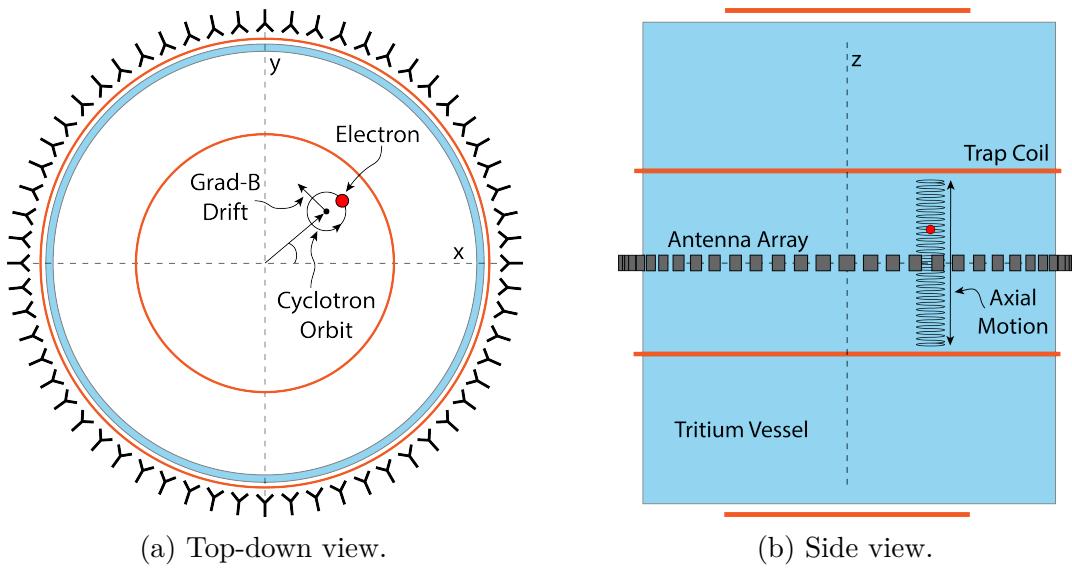


Figure 4.27: An illustration of the conceptual design for an antenna array CRES tritium beta-decay spectrum measurement. The antenna array geometry consists of a 20 cm interior diameter with 60 independent antenna channels arranged evenly around the circumference. The nominal antenna design is sensitive to radiation in the frequency range of 25-26 GHz, which corresponds to the cyclotron frequency of electrons emitted near the tritium beta-spectrum endpoint in a 1 T magnetic field. The array is located at the center of the magnetic trap produced by a set of current-carrying coils. The nominal magnetic trap design is capable of trapping electrons up to 5 cm away from the central axis of the array and traps electrons within an approximately 6 cm long axial region centered on the antenna array.

2545 determined by the size of the background magnetic field. The FSCD design assumes  
 2546 a background magnetic field value of approximately 0.96 T, which results in cyclotron  
 2547 frequencies for electrons with kinetic energies near the tritium beta-spectrum endpoint  
 2548 from 25 to 26 GHz. The component with the next highest frequency is the axial oscillation  
 2549 experienced by electrons with pitch angles of less than  $90^\circ$  [24]. The flat region of the  
 2550 FSCD magnetic trap extends approximately 3 cm above and below the antenna array  
 2551 plane causing electrons to move back and forth as they are reflected from the trap walls.  
 2552 Typical oscillation frequencies are on the order of 10's of MHz, which results in an  
 2553 oscillation period that is  $O(10^3)$  smaller than the duration of a typical CRES event.  
 2554 Therefore, when reconstructing CRES events we treat the electron as occupying only an  
 2555 average axial position at the center of the magnetic trap, since we are not able to resolve  
 2556 the axial position as a function of time. The component of motion with the smallest  
 2557 frequency is  $\nabla B$ -drift caused by radial field gradients in the trap, producing an orbit of  
 2558 the electron around the central axis of the trap with a frequency on the order of a few

2559 kHz, dependent on the pitch angle and the radial position of the electron.

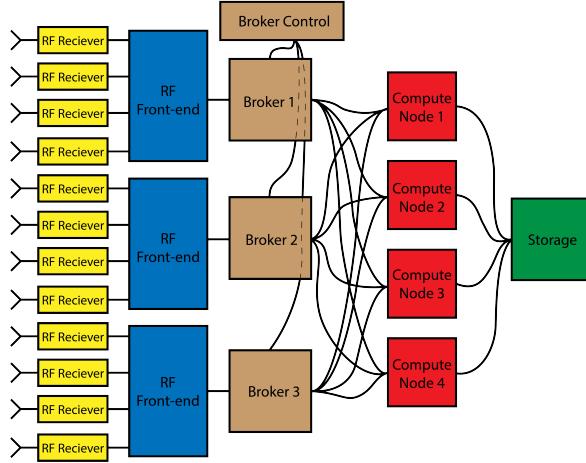


Figure 4.28: A high-level diagram of the DAQ system archctecture envisioned for the FSCD.

2560 The data acquisition (DAQ) system digitizes the signals from the antenna array and  
2561 combines thee data streams into a time-ordered matrix of array snapshots that can be  
2562 used by the reconstruction algorithms. The FSCD DAQ system design [42] is divided into  
2563 three layers 4.28. The first layer is the RF front-end, which includes the antenna array,  
2564 the RF receiver boards, and the digitization electronics. The receiver board contains an  
2565 amplifier, RF mixer, and bandpass filter to enable down-conversion, and is followed by  
2566 the digitization electronics that sample the CRES signals at 200 MHz. In order to achieve  
2567 an adequate signal-to-noise ratio to detect CRES events, the DAQ system for the antenna  
2568 array demonstrator must have a total system noise temperature of  $\approx 10$  K, which we  
2569 can achieve by using low-noise amplifiers and operating at cryogenic temperatures. After  
2570 digitization, the array data must be reorganized from individual data streams sorted  
2571 by channel into array snapshots sorted by time. In order to solve this data transfer  
2572 and networking problem the second layer of the DAQ system consists of a set of broker  
2573 computer nodes that reorganize the array data into time-ordered chunks. This approach  
2574 allows us accommodate different data transfer requirements by scaling the number of  
2575 broker nodes in this layer accordingly. Next, the broker layer distributes these chunks  
2576 of array data to the final layer of the DAQ system, which consists of a set of identical  
2577 reconstruction nodes that perform the calculations required for CRES reconstruction.  
2578 Similar to the broker layer, the number of reconstruction nodes can be increased or  
2579 decreased depending on the amount of computer power required for real-time CRES  
2580 reconstruction.

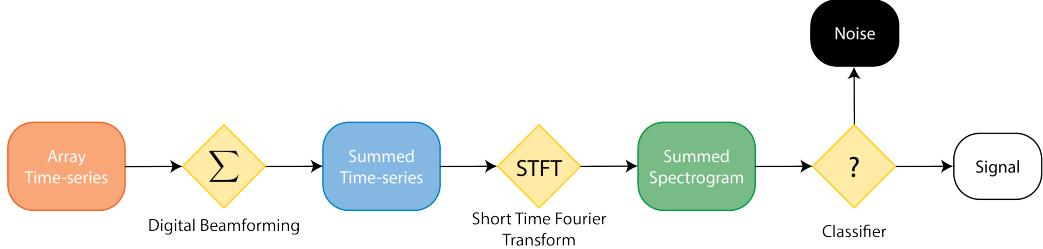


Figure 4.29: A block diagram illustration of the real-time triggering algorithm proposed for antenna array CRES reconstruction.

The design of the FSCD DAQ system is intended to enable a significant portion of the CRES event reconstruction to occur in real-time. The motivation for this comes from the fact that the FSCD antenna array generates approximately 1 exabyte of raw data per year of operation. Therefore, in order to reduce the data-storage requirements, it is ideal to perform at least some of the CRES event reconstruction in real-time so that it is possible to save a reduced form of the data for offline analysis. The first step of the real-time reconstruction would be a real-time signal detection algorithm, which is the focus of this paper. Our approach consists of three main operations performed on the time-series data blocks including digital beamforming, a short time Fourier transform (STFT), and a binary classification algorithm to distinguish between signal and noise data (see Figure 4.29).

#### 4.4.2.2 Real-time Signal Detection

The first step in the real-time detection algorithm is digital beamforming, which is a phased summation of the signals received by individual antennas in the array (see Figure 5.21). The phase shifts correspond to the path length differences between a spatial position and each individual antenna such that, when there is an electron located at the beamforming position, all the signals received by the array constructively interfere. Since we do not know ahead of time where an electron will be produced in the detector, we define a grid of beamforming positions that cover the entire region where electrons can be trapped and perform a phased summation for each of these points for every time-step in the array data block. As we saw in Section 4.4.2.1, the axial oscillation of the electrons prevents us from resolving its position along the Z-axis of the trap, therefore our beamforming grid need only cover the possible positions of the electron in the two-dimensional plane defined by the antenna array.

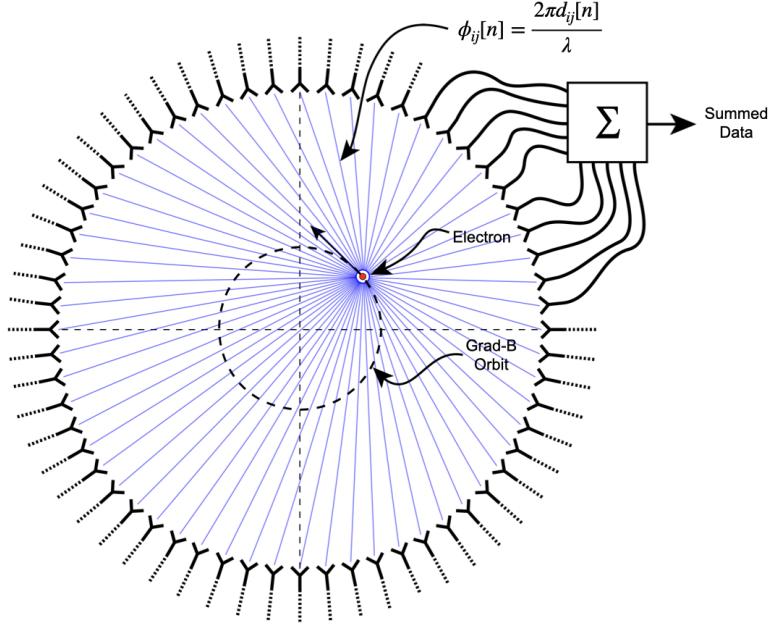


Figure 4.30: An illustration of the digital beamforming procedure. The blue lines indicate the various distances from the beamforming position to the antenna. In the situation depicted the actual position of the electron matches the beamforming position, so we should expect constructive interference when the phase shifted signals are summed. To prevent the electron's  $\nabla B$ -motion from moving the electron off of the beamforming position, the beamforming phase include a time-dependence to follow the trajectory of the electron in the magnetic trap.

2605 The equation defining digital beamforming can be expressed as

$$\mathbf{y}[n] = \Phi^T[n]\mathbf{x}[n], \quad (4.39)$$

2606 where  $\mathbf{x}[n]$  is array snapshot vector at the sampled time  $n$ ,  $\Phi[n]$  is the matrix of  
 2607 beamforming phase shifts, and  $\mathbf{y}[n]$  is summed output vector that contains the voltages  
 2608 for each of the summed channels that correspond to a particular beamforming position.  
 2609 The elements of the beamforming phase shift matrix can be expressed as a weighted  
 2610 complex exponential

$$\Phi_{ij}[n] = A_{ij}[n] \exp(2\pi i \phi_{ij}[n]), \quad (4.40)$$

2611 where the indices  $i$  and  $j$  label the beamforming and antenna positions respectively. The  
 2612 weight  $A_{ij}$  accounts for the relative power increase for antennas that are closer to the  
 2613 position of the electron, and  $\phi_{ij}$  is the total beamforming phase shift for the  $j$ -th antenna  
 2614 at the  $i$ -th beamforming position.

2615 The beamforming phase shift is a sum of two terms

$$\phi_{ij}[n] = \frac{2\pi d_{ij}[n]}{\lambda} + \theta_{ij}[n], \quad (4.41)$$

2616 where the first term is the phase shift originating from the path length difference ( $d_{ij}[n]$ )  
2617 between the beamforming and antenna positions, which are represented by the vectors  
2618 ( $r_j, \theta_j$ ) and ( $r_i, \theta_i[n]$ ), and the second term is the angular separation ( $\theta_{ij}[n]$ ) of the two  
2619 positions. The angular separation enters into the beamforming phase due to an effect  
2620 caused by the circular orbit of the electron that produces radiation whose phase is linearly  
2621 dependent on the relative azimuthal position of the antenna [43,44]. The time-dependence  
2622 of the beamforming phases is intended to correct for the effects of  $\nabla B$ -drift, which cause  
2623 the guiding centers of electrons to orbit the center of the magnetic trap. By including a  
2624 linear time-dependence in the azimuthal beamforming position,

$$\theta_i[n] = \omega_{\nabla B} t[n] + \theta_{i,0}, \quad (4.42)$$

2625 where  $\omega_{\nabla B}$  is the azimuthal grad-B drift frequency,  $t[n]$  is the time vector and,  $\theta_{i,0}$  is the  
2626 starting azimuthal position, we can configure the beamforming phases to effectively track  
2627 the XY-position of the guiding center over the event duration. Predicting accurate values  
2628 of  $\omega_{\nabla B}$  for a specific trap and set of kinematic parameters will be done by simulations,  
2629 which are performed using the Kassiopeia software package [23] by Project 8.

2630 After digital beamforming, we apply a short-time Fourier transform (STFT) to the  
2631 summed time-series to obtain the frequency spectrum representation of the signals (see  
2632 Figure 4.31). From the detection perspective, the frequency representation of the CRES  
2633 data is advantageous compared to the time domain, because the frequency spectra of  
2634 CRES signals are well-approximated by a frequency and amplitude modulated sinusoidal  
2635 whose carrier frequency increases as a linear chirp. The modulation is caused by the axial  
2636 oscillation of the electron in the magnetic trap and produce frequency spectra that are  
2637 well-described by a small number of frequency components. The linear chirp is caused  
2638 by the energy loss due to cyclotron radiation, which results in a relatively slow increase  
2639 in the frequency components of the CRES signal over time. During the standard Fourier  
2640 analysis window for the FSCD of 40.96  $\mu$ sec, we expect a typical CRES signal to increase  
2641 in frequency by approximately 15 kHz, which is smaller than the frequency bin width  
2642 given the 200 MHz sample rate. Therefore when considering a single frequency spectrum  
2643 it is justifiable to neglect the effects of the linear frequency chirp.

2644 In the cases where the electron's pitch angle is  $\gtrsim 88.5^\circ$ , the majority of the signal

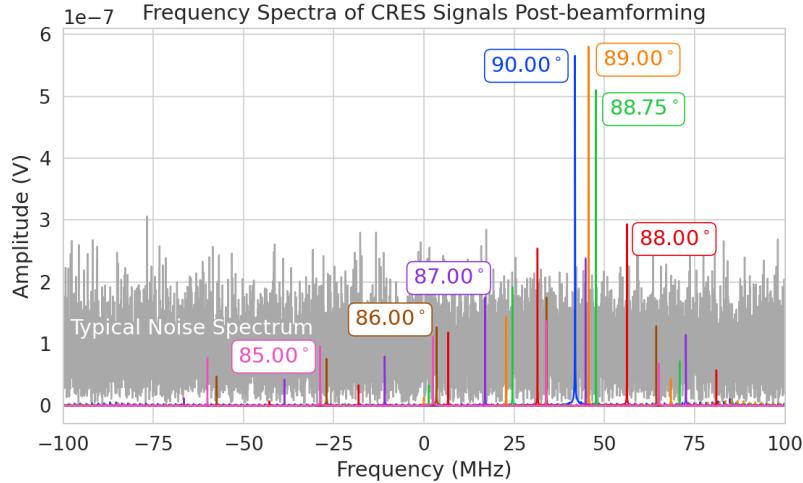


Figure 4.31: Frequency spectra of simulated CRES signals post-beamforming. The signal of a  $90^\circ$  electron consists of a single frequency component that is easy to detect with a power threshold on the frequency spectrum. This power threshold is still effective for signals with relatively large pitch angles such as  $89.0^\circ$  and  $88.75^\circ$ , which are composed of a main carrier and a few small sidebands. Signals with smaller pitch angles, below about  $88.5^\circ$ , tend to be dominated by sidebands such that no single frequency component can be reliably distinguished from the noise with a power threshold.

power is contained in a single frequency component, with the remaining signal power contained in a small number of sidebands proportional to the electron's axial modulation (see Figure 4.31). In these cases detection is relatively straight-forward by implementing a power threshold on the STFT, since the amplitude of the main signal peak is distinct from the thermal noise spectrum. However, as the pitch angle of the electron is decreased below  $88.5^\circ$ , the modulation index of the signal increases causing the maximum amplitude of the frequency spectrum to be comparable to typical noise fluctuations. At this point, the power threshold trigger is no longer able to distinguish between signal and noise leading to a reduction in detection efficiency. The neutrino mass sensitivity of the FSCD is directly linked to the overall detection efficiency. And, because the distribution of electron pitch angles is effectively uniformly distributed across the range of pitch angles that can be trapped, the overall detection efficiency is directly influenced by the range of pitch angles that have detectable signals. Therefore, utilizing a signal detection algorithm that can more effectively identify signals with pitch angles less than  $88.5^\circ$  will improve both detection efficiency and ultimately the neutrino mass sensitivity of the FSCD and other CRES experiments.

Modeling the detection performance of alternative signal detection algorithms for

2662 the FSCD requires that we pose the signal detection problem in a consistent manner.  
2663 The approach we take is to perform a binary hypothesis test on the frequency spectra  
2664 generated by the STFT. Mathematically, this is expressed as,

$$\mathcal{H}_0 : y[n] = \nu[n] \quad (4.43)$$

$$\mathcal{H}_1 : y[n] = x[n] + \nu[n]. \quad (4.44)$$

2665 Where under hypothesis  $\mathcal{H}_0$ , the vector representing the frequency spectrum ( $y[n]$ ) is  
2666 composed of pure white Gaussian noise (WGN) represented by  $\nu[n]$ , and under hypothesis  
2667  $\mathcal{H}_1$  the frequency spectrum is composed of a CRES signal ( $x[n]$ ) with added WGN. The  
2668 dominant source of noise in a FSCD-like experiment is expected to be thermal Nyquist-  
2669 Johnson noise, which is well approximated by a WGN distribution. In order to decide  
2670 between these two hypotheses we follow the standard Neyman-Pearson approach by  
2671 performing a log-likelihood ratio test between the probability distributions of the signal  
2672 classifier output under  $\mathcal{H}_1$  and  $\mathcal{H}_0$  [34]. The output of the log-likelihood ratio test is  
2673 called the test statistic, which is used to assign the data as belonging to the noise ( $\mathcal{H}_0$ )  
2674 or signal ( $\mathcal{H}_1$ ) classes by setting a decision threshold on the value of the test statistic.

2675 In practice, we select the decision threshold by finding the value of the test statistic  
2676 that guarantees an acceptable rate of false positives and then attempt to maximize  
2677 the signal detection probability under that fixed false positive rate. Because the signal  
2678 classifier will be used to evaluate the spectra of  $O(10^2)$  beamforming positions every  
2679  $40.96 \mu\text{sec}$ , we will require the signal classifiers to operate with decision thresholds that  
2680 provide false positive rates significantly smaller than 1%. This reduces the burden placed  
2681 on later stages of the CRES reconstruction chain to reject these false positives and  
2682 decreases the overall likelihood of reconstructing a false event. Below, we calculate the  
2683 probability distributions that allow us characterize how different detection algorithms  
2684 will perform for CRES signals in an FSCD experiment.

### 2685 **4.4.3 Signal Detection Algorithms**

#### 2686 **4.4.3.1 Power Threshold**

2687 The power threshold detection algorithm uses the maximum amplitude of the frequency  
2688 spectra as the detection test statistic. To model the performance of this approach,  
2689 consider first the case where the signal is pure WGN. For a single bin in the frequency  
2690 spectrum, the probability that the amplitude falls below a specific threshold value is

2691 given by the Rayleigh cumulative distribution function (CDF),

$$\text{Ray}(x; \tau) = 1 - \exp(-|x|^2/\tau), \quad (4.45)$$

2692 where the complex amplitude of the frequency bin is  $x$ , and  $\tau$  is the WGN variance.  
 2693 Because the noise samples for each frequency bin are independent and identically dis-  
 2694 tributed (IID), the probability that every bin in the frequency spectrum falls below the  
 2695 threshold is the joint CDF formed by the product of each individual frequency bin CDF,

$$F_0(x; \tau, N_{\text{bin}}) = \text{Ray}(x; \tau)^{N_{\text{bin}}}. \quad (4.46)$$

2696 The PDF for the power threshold classifier can then be obtained by differentiating the  
 2697 CDF.

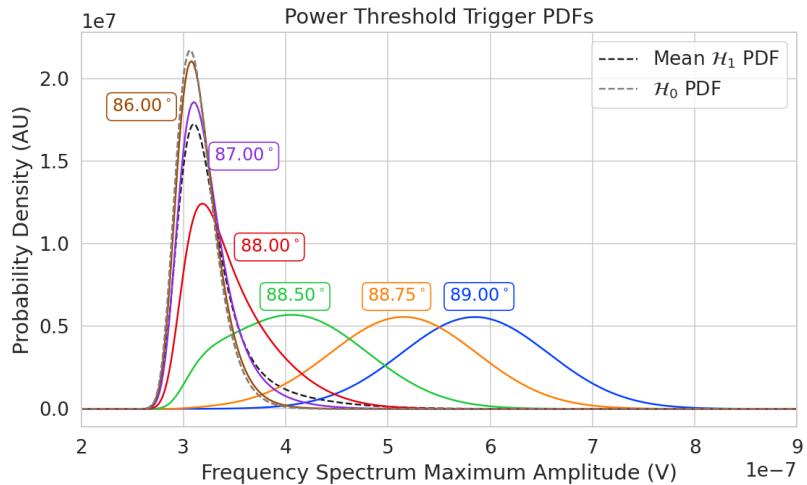


Figure 4.32: PDFs of the power threshold test statistic for CREs signals with various pitch angles as well as the PDF for the noise-only signal case. The average PDF computed for pitch angles ranging from 85.5 to 88.5° is also shown. As the pitch angle is decreased the signal PDF converges towards the noise PDF which indicates that the power threshold trigger is unable to distinguish between small pitch angle signals and noise.

2698 The probability distribution for the power threshold classifier under  $\mathcal{H}_1$  is formed in  
 2699 a similar way, but the frequency bins that contain signal must be treated separately. For  
 2700 a frequency bin that contains both signal and noise we can describe the probability that  
 2701 the amplitude of the bin will fall below our threshold using the Rician CDF,

$$\text{Rice}(x; \tau, \nu) = 1 - Q_1 \left( \frac{|\nu|}{\sqrt{2\tau}}, \frac{|x|}{\sqrt{2\tau}} \right), \quad (4.47)$$

2702 where the parameter  $|\nu|$  defines the noise-free amplitude of the signal and  $Q_1$  is the  
 2703 Marcum Q-function. This time the CDF that describes the probability that the entire  
 2704 spectrum falls below the decision threshold is the product of both signal and noise CDFs,

$$F_1(x; \tau, \nu, N_{\text{bin}}, N_s) = \text{Ray}(x; \tau)^{N_{\text{bin}} - N_s} \prod_{k=0}^{N_s} \text{Rice}(x; \tau, \nu_k). \quad (4.48)$$

2705 The first half of Equation 4.48 is the contribution from the bins in the frequency spectrum  
 2706 that contain only noise, and the second half is the product of the Rician CDFs for the  
 2707 frequency bins that contain signal peaks with a noise-free amplitude of  $|\nu_k|$ . In Figure  
 2708 4.32 we show plots of example PDFs under  $\mathcal{H}_1$  and  $\mathcal{H}_0$ .

#### 2709 4.4.3.2 Matched Filtering

2710 The shape of a CRES signal is completely determined by the initial conditions of the  
 2711 electron as it is emitted from beta-decay, which implies that it is possible to apply  
 2712 matched filtering as a signal detection algorithm. With a matched filter one uses the  
 2713 shape of the known signal, which is called a template, to filter the incoming data by  
 2714 computing the convolution between the signal and the data [34]. For cases where the  
 2715 signal is buried in WGN, the matched filter is the optimal detector in that it achieves  
 2716 the maximum probability of a true detection for a fixed false positive rate. Since CRES  
 2717 signals have an unknown shape but are deterministic, we can apply a matched filter by  
 2718 using simulations to generate a large number of signal templates called a template bank,  
 2719 which spans the parameter space of possible signals. Then at detection time, we use the  
 2720 template bank to identify signals by performing the matched filter convolution for each  
 2721 template in an exhaustive search.

2722 As we saw from the frequency spectra in Figure 4.31, CRES signals are highly periodic  
 2723 in nature. In such cases, it is advantageous to utilize the convolution theorem to replace  
 2724 the matched filter convolution with an inner product in the frequency-domain. With the  
 2725 convolution theorem, the matched filter test statistic that describes the detection of a  
 2726 signal buried in WGN using a matched filter template bank is given by

$$\mathcal{T} = \max_{\mathbf{h}} \left| \sum_{n=0}^{N_{\text{bin}}} h^\dagger[n] y[n] \right|, \quad (4.49)$$

2727 where  $h^\dagger[n]$  is the complex conjugate of the signal template. For the case when our  
 2728 template bank consists of only a single template it is possible to derive an exact analytical

2729 form for the PDF describing the matched filter test statistic. First, we derive PDF under  
 2730 the signal hypothesis, where the equation describing the matched filter test statistic, also  
 2731 known as the matched filter score, becomes

$$\mathcal{T} = \left| \sum_{n=0}^{N_{\text{bin}}} h^\dagger[n] y[n] \right|. \quad (4.50)$$

2732 Each noisy frequency bin represented by  $y[n]$  is the sum between value of the signal  
 2733 at that bin and complex WGN, which means that  $y[n]$  is itself Gaussian distributed.  
 2734 Therefore, the value of the inner product between the template and the data is also a  
 2735 complex Gaussian variable; and, since the matched filter score is the magnitude of this  
 2736 inner product, it must follow a Rician distribution.

2737 We can derive the equation for the Rician PDF by expressing the matched filter  
 2738 template  $\mathbf{h}$  in terms of the corresponding simulated signal, which we write as  $\mathbf{x}_h$  to  
 2739 distinguish from the signal in the data. Using the standard normalization and assuming  
 2740 uncorrelated WGN, the matched filter templates can be written as

$$\mathbf{h} = \frac{\mathbf{x}_h}{\sqrt{\tau |\mathbf{x}_h|^2}} \quad (4.51)$$

2741 where  $\tau$  is the noise variance. Inserting this into Equation 4.49 and expressing the data  
 2742 as a sum between a signal and a WGN vector yields,

$$\mathcal{T} = \frac{1}{\sqrt{\tau |\mathbf{x}_h|^2}} \left| \sum_{n=1}^{N_{\text{bin}}} x_h[n] (x[n] + \nu[n]) \right|. \quad (4.52)$$

2743 Next, we transform the expression by isolating the randomly distributed components  
 2744 giving

$$\mathcal{T} = \frac{\left| \sum_{n=1}^{N_{\text{bin}}} x_h[n] x[n] \right|}{\sqrt{\tau |\mathbf{x}_h|^2}} + \frac{1}{\sqrt{\tau |\mathbf{x}_h|^2}} \left| \sum_{n=1}^{N_{\text{bin}}} x_h[n] \nu[n] \right|. \quad (4.53)$$

2745 The first term of 4.53 can be simplified by using the Cauchy-Schawrz inequality to express  
 2746 the magnitude of the inner product in terms of the magnitudes of the signal and template  
 2747 as well as an orthogonality constant which we call "match" ( $\Gamma$ ). Using this we obtain,

$$\mathcal{T} = |\mathbf{h}| |\mathbf{x}| \Gamma + \frac{1}{\sqrt{\tau |\mathbf{x}_h|^2}} \left| \sum_{n=1}^{N_{\text{bin}}} x_h[n] \nu[n] \right|. \quad (4.54)$$

2748 The second term is a sum of Gaussian distributed variables, which we should expect also  
2749 follows a Gaussian distribution. Each of the samples  $\nu[n]$  is described by

$$\nu[n] \sim \mathcal{N}(0, \tau), \quad (4.55)$$

2750 where  $\mathcal{N}(0, \tau)$  is a complex Gaussian distribution with zero mean and variance  $\tau$ . There-  
2751 fore,

$$\frac{x_h[n]}{\sqrt{\tau|\mathbf{x}_h|^2}} \nu[n] \sim \mathcal{N}\left(0, \frac{x_h[n]^2}{|\mathbf{x}_h|^2}\right), \quad (4.56)$$

$$\sum_{n=1}^{N_{\text{bin}}} \frac{x_h[n]}{\sqrt{\tau|\mathbf{x}_h|^2}} \nu[n] \sim \mathcal{N}\left(0, \frac{\sum_{n=1}^{N_{\text{bin}}} x_h[n]^2}{|\mathbf{x}_h|^2}\right) = \mathcal{N}(0, 1), \quad (4.57)$$

$$|\mathbf{h}| |\mathbf{x}| \Gamma + \sum_{n=1}^{N_{\text{bin}}} \frac{x_h[n]}{\sqrt{\tau|\mathbf{x}_h|^2}} \nu[n] \sim \mathcal{N}(|\mathbf{h}| |\mathbf{x}| \Gamma, 1). \quad (4.58)$$

2752 We see that  $\mathcal{T}$  is magnitude of a complex variable with mean  $|\mathbf{h}| |\mathbf{x}| \Gamma$  and variance one. In  
2753 order to simply the expression a bit further, we define the quantity  $\mathcal{T}_{\text{ideal}} = |\mathbf{h}| |\mathbf{x}| \Gamma$ , which  
2754 we call the ideal matched filter score, because it represents the value of the matched  
2755 filter inner product that we would expect if no noise was present in the signal. We can  
2756 write the matched filter test statistic as the magnitude of a two-dimensional vector in  
2757 the complex plane

$$\mathcal{T} = |(\mathcal{T}_{\text{ideal}} + n_r, n_i)|, \quad (4.59)$$

2758 where  $n_r$  and  $n_i$  are the real and imaginary components of the noise each with variance  
2759  $1/2$ , which is modeled by a Rician distribution with shape factor  $\mathcal{T}_{\text{ideal}}$ . Therefore, the  
2760 probability distribution of the matched filter test statistic is given by,

$$P_1(x; \mathcal{T}_{\text{ideal}}) = 2x \exp(- (x^2 + \mathcal{T}_{\text{ideal}}^2)) I_0(2x\mathcal{T}_{\text{ideal}}), \quad (4.60)$$

2761 where  $I_0$  is the zeroth-order modified Bessel function.

2762 The shape of the matched filter score distribution is controlled by the parameter  
2763  $\mathcal{T}_{\text{ideal}}$ , which is effectively the value of the matched filter score if the data contained no  
2764 noise. Without noise, the data vector reduces to the signal,  $\mathbf{x}$ , in which case Equation  
2765 4.50 becomes the magnitude of an inner product between two vectors. We can write  
2766 the magnitude of an inner product in terms of the lengths of the individual vectors and  
2767 a constant that describes the degree of orthogonality between them. Applying this to

2768 Equation 4.50, we obtain

$$\mathcal{T}_{\text{ideal}} = |\mathbf{h}^\dagger \cdot \mathbf{x}| = |\mathbf{h}| |\mathbf{x}| \Gamma \quad (4.61)$$

2769 where  $\Gamma$  describes the orthogonality between  $\mathbf{h}$  and  $\mathbf{x}$ . From the point of view of matched  
2770 filtering, we can interpret  $\Gamma$  as describing how well the template matches the underlying  
2771 signal in the data.

2772 The matched filter score PDF under the noise hypothesis can be readily obtained  
2773 from Equation 4.60 by setting the value of  $\mathcal{T}_{\text{ideal}}$  to zero, since the data contains no signal  
2774 in the noise case. Doing this, we obtain the Rayleigh distribution that describes the  
2775 matched filter score under  $\mathcal{H}_0$ ,

$$P_0(x) = 2x \exp(-x^2). \quad (4.62)$$

2776 Equations 4.60 and 4.62 describe the behavior of the matched filter test statistic  
2777 under  $\mathcal{H}_0$  and  $\mathcal{H}_1$  for a single template. However, defining a PDF that describes the  
2778 matched filter test statistic in the case of multiple templates is in general a mathematically  
2779 intractable problem, since there is no guarantee of orthogonality between matched filter  
2780 templates. This leads to correlations between the matched filter scores of different  
2781 templates because only one sample of noise is used to compute the matched filter scores  
2782 of the template bank. In order to proceed, we need to make the simplifying assumption  
2783 that we can treat the matched filter scores as IID variables, which allows to ignore  
2784 correlations between templates. The overall effect of this will be an underestimate of the  
2785 performance of the matched filter, since we are under counting the number of templates  
2786 that could contribute a detectable score.

2787 For  $\mathcal{H}_0$  we model the probability that the matched filter score falls below our threshold  
2788 using the CDF obtained by integrating Equation 4.62. Because we are assuming that  
2789 the matched filter scores using different templates are independent, the probability that  
2790 the matched filter score for all templates falls below a threshold value is the joint CDF  
2791 formed by multiplying the CDF for each template. Under  $\mathcal{H}_0$  this is

$$F_0(x) = \left(1 - e^{-x^2}\right)^{N_t}, \quad (4.63)$$

2792 where  $x$  is the matched filter score threshold and  $N_t$  is the number of templates. We  
2793 should expect that the distribution describing the matched filter template bank maximum  
2794 score depends on  $N_t$ , because with more templates there is a greater chance of a random  
2795 match between the template and data.

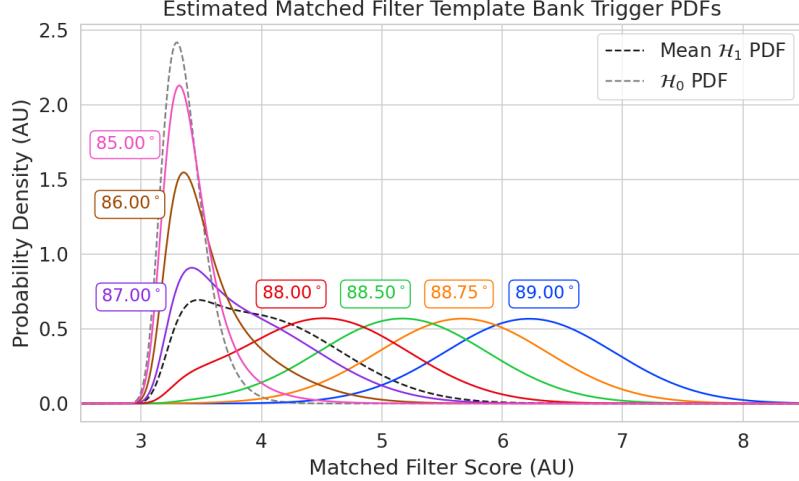


Figure 4.33: Plots of the estimated PDFs for the matched filter template bank test statistic for CRES signals with various pitch angles as well as the estimated PDF for the noise only signal case. We assume an estimated number of templates of  $10^5$  and perfect match between signal and template i.e.  $\Gamma_{\text{best}} = 1$ . The mean PDF includes signals ranging from  $85.5 - 88.5^\circ$  in pitch angle. There is a much larger distinction between the signal PDFs at small pitch angle compared to the power threshold indicating a higher detection efficiency for these signals.

For  $\mathcal{H}_1$ , we start by denoting the CDF of the best matching template as  $F_{\text{best}}(x; \mathcal{T}_{\text{best}})$ , and treat the matched filter scores for all other templates as negligible ( $\mathcal{T}_{\text{ideal}} \approx 0$ ). Then we form the joint CDF by combining the distributions for all templates used during detection. Since we are exhaustively checking the matched filter scores, the number of templates checked will be a randomly distributed variable that ranges from zero to the total number of available templates. If we assume that signals are uniformly distributed across the parameter space spanned by the template bank then on average we check  $(N_t - 1)/2 \approx N_t/2$  templates for each inference. Therefore, the estimated CDF under  $\mathcal{H}_1$  is

$$F_1(x; \mathcal{T}_{\text{best}}) = F_{\text{best}}(x; \mathcal{T}_{\text{best}}) \left(1 - e^{-x^2}\right)^{N_t/2}. \quad (4.64)$$

In Figure 4.33 we show plots of the estimated matched filter template bank classifier PDFs under both  $\mathcal{H}_0$  and  $\mathcal{H}_1$ .

#### 4.4.3.3 Machine Learning

In this paper we focus on Convolutional Neural Networks (CNN) as an example of a machine learning based signal classifier. CNNs are constructed using a series of

convolutional layers, each composed of a set of filters that are convolved with the input data. The individual convolutional filters can be viewed as matched filter templates that are learned from a set of simulated data rather than being directly generated. This opens the possibility of finding a more efficient representation of the matched filter templates during the training process that can potentially reduce computational cost at inference time while still offering good classification performance.

The machine learning approach is distinct from both the power threshold and matched filtering in that we do not attempt to manually engineer a test statistic that is computed from the data for classification. Instead, we attempt calculate the test statistic by constructing a differentiable function that maps the complex frequency series generated by the STFT to a binary classification as either signal or noise. The test statistic for the machine learning classifier can be expressed as

$$\mathcal{T} = G(\mathbf{y}; \boldsymbol{\Omega}) \quad (4.65)$$

where  $\mathbf{y}$  is the noisy data vector and  $G(\mathbf{y}; \boldsymbol{\Omega})$  is the machine learning model parameterized by the weights  $\boldsymbol{\Omega}$ . By using supervised learning on a labeled set of training signals, we can modify the function parameters to learn the mapping from the data to the likelihood of  $\mathbf{y}$  belonging to either  $\mathcal{H}_1$  or  $\mathcal{H}_0$ .

Table 4.1: A summary of the CNN model layers and parameters. The output of each 1D-Convolution and Fully Connected layer is passed through a LeakyReLU activation function and re-normalized using batch normalization before being passed to the next layer in the model. The output of the final Fully Connected layer in the model is left without activation so that the model outputs can be directly passed to the Binary Cross-entropy loss function used during training.

Layer	Type	Input Channels	Output Channels	Parameters
1	1D-Convolution	2	15	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 1$ )
2	Maximum Pooling	15	15	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 4$ )
3	1D-Convolution	15	20	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 1$ )
4	Maximum Pooling	20	20	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 4$ )
5	1D-Convolution	20	25	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 1$ )
6	Maximum Pooling	25	25	( $N_{\text{kernel}} = 4$ , $N_{\text{stride}} = 4$ )
7	Fully Connected	3200	512	NA
8	Fully Connected	512	64	NA
9	Fully Connected	64	2	NA

The CNN architecture used for this work is summarized by Table 4.1. No strategic hyper-parameter optimization approach was implemented beyond the manual testing

of different CNN architecture variations, so this particular model is best viewed as a proof-of-concept rather than a rigorously optimized design. Numerous model variations were tested, some with significantly more layers and convolutions filters per layer, as well as others that were even smaller than the architecture in Table 4.1. Ultimately, the model architecture choice was driven by the motivation to find the minimal model whose classification performance was still comparable to the larger CNN’s tested, because of the importance of minimizing computational cost in real-time applications. It is possible that more sophisticated machine learning models could improve upon the classification results achieved here, but we leave this investigation for future work.

#### 4.4.4 Methods

##### 4.4.4.1 Data Generation

To test the triggering performance of the classifiers, simulated CRES signals were generated using the Locust simulations package [25, 43] developed by the Project 8 collaboration. Locust uses the separately developed Kassiopeia package to calculate the magnetic fields produced by a user defined set of current carrying coils along with any specified background magnetic fields, resulting in a magnetic trap. Next, Kassiopeia calculates the trajectory of an electron in this magnetic field starting from a set of user specified initial conditions. The Locust software then uses the electron trajectories from Kassiopeia to calculate the resulting electromagnetic fields using the Liénard-Wiechert equations, and determine the voltages generated in the antenna array with the antenna transfer function. Locust then simulates the down-conversion, filtering, and digitization steps resulting in the simulated CRES signals for an electron.

The shape of the received CRES signal is determined by the initial kinematic parameters, including the starting position of the electron, the starting kinetic energy of the electron, and the pitch angle. For the studies performed here we constrain ourselves to a single initial electron position located at  $(x, y, z) = (5, 0, 0)$  mm, and using this starting position we generate two datasets by varying the initial kinetic energy and the starting pitch angle. The first dataset consists of a two-dimensional square grid of kinetic energy and pitch angle spanning an energy range from 18575-18580 eV with a spacing of 0.1 eV, and pitch angles from 85.5-88.5° with a spacing of 0.001°, resulting in 153051 signals with a unique energy-pitch angle combination. This dataset is intended to represent a matched filter template bank. The second dataset was generated by randomly sampling uniform probability distributions covering the same parameter space to produce approximately

2861 50000 signals randomly parameterized in energy and pitch angle. This dataset provides  
2862 the training and test data for the machine learning approach, and acts as a representative  
2863 sample of signals to evaluate the performance of the matched filter template bank.

2864 Each signal was simulated for a duration of  $40.96 \mu\text{s}$ , which is equivalent to 8192  
2865 samples at the FSCD digitization rate, and begins at time  $t = 0 \text{ s}$  for all simulations.  
2866 This duration represents a single frequency spectrum generated by the STFT. The output  
2867 of the Locust simulation is a matrix of array snapshots with size given by the number of  
2868 channels times the event length ( $N_{\text{ch}} \times N_{\text{sample}}$ ), which we pre-process using the digital  
2869 beamforming summation and STFT described in Section 4.4.2.2. The  $\nabla B$ -drift correction  
2870 uses the exact value of  $\omega_{\nabla B}$ , obtained from the Kassiopeia simulation of that electron.  
2871 In practice, an average value for  $\omega_{\nabla B}$  could be used, because there is limited variation in  
2872 drift frequency across this parameter space.

#### 2873 4.4.4.2 Template Number and Match Estimation

2874 The estimated PDF for the matched filter template bank depends on the score of the  
2875 best matching template or equivalently the match of the best template ( $\Gamma_{\text{best}}$ ) as well  
2876 as the number of templates. One expects that with a higher number of templates the  
2877 average value of  $\Gamma_{\text{best}}$  will increase, however, there is a point of diminishing returns at  
2878 which more templates will not significantly increase match, but will still increase the  
2879 likelihood of false positives. Therefore, it is desirable to use the minimum number of  
2880 templates that provide an acceptable mean value of  $\Gamma_{\text{best}}$ .

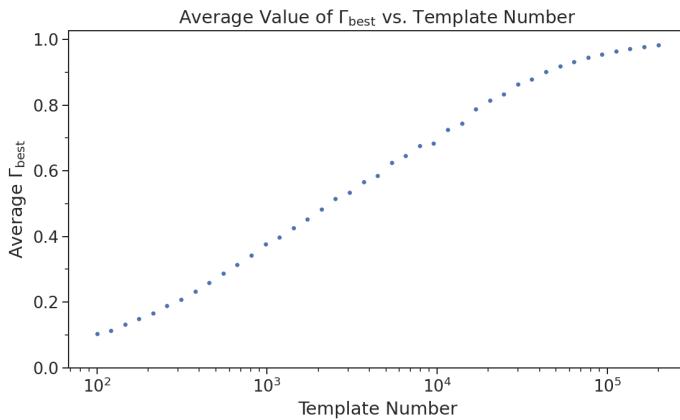


Figure 4.34: The mean match of the matched filter template bank to a test set of randomly parameterized signals as a function of the number or density of templates. The parameter space includes pitch angles from  $85.5 - 88.5^\circ$  and energies from 18575 – 18580 eV.

2880

2881 To quantify the relationship between match and template number, we calculated  
2882 the mean match of the random dataset to a selection of templates obtained from the  
2883 regularly spaced dataset. The results are shown in Figure 4.34, where we find that the  
2884 average value of  $\Gamma_{\text{best}}$  is an exponential function of the number of templates. From this  
2885 plot we select the desired value of mean match at which we would like to evaluate the  
2886 matched filter PDF and can infer the required number of templates.

#### 2887 4.4.4.3 CNN Training and Data Augmentation

2888 To prepare the data for training the model, we split the random dataset in half to create  
2889 distinct training and test datasets. Additionally, a randomly selected 20% of the training  
2890 data is isolated for use as a validation set during the training loop. The size of the  
2891 training, validation, and test datasets are then tripled by appending two additional copies  
2892 of the data to increase the sample size of the dataset after data augmentation. The  
2893 data is loaded with no noise, which is added to each data batch during the training  
2894 phase by generating a new noise sample from a complex WGN distribution. In order to  
2895 ensure an even split between signal and noise data we append to the noise-free signals an  
2896 equal number of empty signals composed of all zeros. Therefore, as the data is randomly  
2897 shuffled during training, on average an equal number of empty signals will be included  
2898 with the training signals. After adding the sample of WGN to the data batch, the empty  
2899 signals represent the noise-only data that the model must distinguish from signal data.

2900 As the training signals are loaded we apply a unique random phase shift as the  
2901 first form of data augmentation. Since the data is generated using the same initial  
2902 axial position and cyclotron orbit phase, the randomization is an attempt to prevent  
2903 overtraining on these features. During each training epoch the data is randomly shuffled  
2904 and split into batches of 2500 signals. Each batch of signals is then circularly shifted  
2905 by a random number of frequency bins to simulate a kinetic energy shift from  $-75$  to  
2906  $20$  eV to simulate a training dataset with a larger energy range. Next, a sample of  
2907 complex WGN, consistent with the expected 10 K Nyquist-Johnson noise expected for  
2908 the FSCD, is generated and added to the signal, which prevents overtraining on noise  
2909 features. As a final step, the data is renormalized by the standard deviation of the noise  
2910 so that the range of values in the data is close to  $[-1, 1]$ , which helps ensure well-behaved  
2911 back-propagation.

2912 The Binary Cross-entropy loss function is used to compute the loss for each batch of  
2913 data and the model weights are updated using the ADAM optimizer with a learning rate  
2914 of  $5 \times 10^{-3}$ . After each training epoch, the loss and classification accuracy of the validation

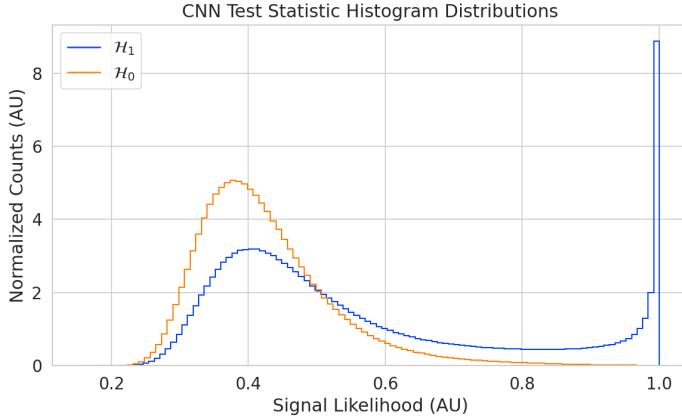


Figure 4.35: Histograms of the trained CNN model output from the test dataset. The blue histogram shows the model outputs for signal data. The oddly shaped peak near the end is the result of the softmax function mapping the long tail of the raw output distribution to the range [0, 1].

dataset are computed to monitor for overtraining. It was noticed that the relatively high noise power and the fact that a new sample of noise was used for each batch together provided a strong form of regularization, since no evidence of over-training was observed even after several thousand epochs. Typically, the loss and classification accuracy of the model converged after a few hundred training epochs, but the training loop was extended to 3000 epochs to attempt to achieve the best possible performance. The training procedure generally took about 24 hrs using a single NVIDIA V100 GPU [45].

After training the model, we use it to classify the test dataset and generate histograms of the model outputs for both classes of data. The data augmentation procedure for the evaluation of the test data mirrors the training procedure without the validation split. Since a random circular shift and a new sample of WGN is added to each batch, the testing evaluation loop is run for 100 epochs to get a representative sample of noise and circular shifts. The model outputs for each batch are passed through a softmax activation and then combined into histograms, which we show in Figure 4.35.

## 4.4.5 Results and Discussion

### 4.4.5.1 Trigger Classification Performance

Using the matched filter and power threshold CDFs, along with the classification results from the CNN, we compare detection performance by computing receiver operating characteristic (ROC) curves. Specifically, we compare the detection performance averaged

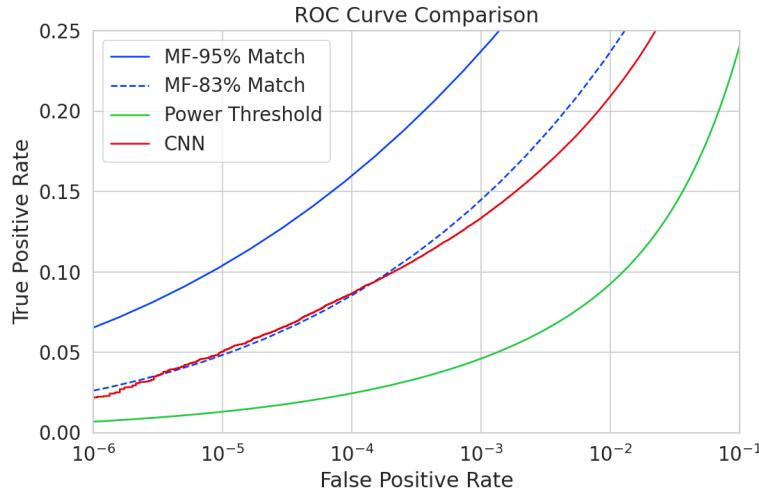


Figure 4.36: ROC curves describing the detection efficiency or true positive rates for the three signal classification algorithms examined in this paper.

over the full signal parameter space in order to get a measure of the overall detection  
 efficiency achieved by each algorithm. For the power threshold and matched filter  
 algorithms, we obtain the mean ROC curve by taking the average over all signals in the  
 regularly spaced dataset. In the case of the matched filter, we examine two cases using  
 different numbers of templates, which have different values of mean match. The ROC  
 curve describing the CNN is obtained by forming a histogram of the network outputs  
 for each class of signal and from this computing the estimated CDFs and ROC curve.  
 In Figure 4.36, we show the ROC curves obtained for each of the detection algorithms,  
 visualized in terms of true positive rate and false positive rate.

The true positive rate of a signal classifier is equivalent to its detection efficiency, and  
 we see that for the population of signals with pitch angles  $< 88.5^\circ$  the power threshold  
 has a consistently lower detection efficiency than the CNN and the matched filter. This  
 result could have been predicted from the visualization of signal spectra in Figure 4.31,  
 where we see that there is no way to distinguish between a noise peak and a signal peak  
 with high confidence at small pitch angles. The CNN offers a significant and consistent  
 increase in detection efficiency over the power threshold approach, with the relative  
 improvement in detection efficiency increasing as the false positive rate decreases. If  
 we compare the CNN to the matched filter, we see that the performance of the tested  
 network is roughly equivalent to a matched filter detector with an average match of  
 about 83%, which uses approximately 20000 matched filter templates. The overall best  
 detection efficiency is achieved by the matched filter classifier if a large enough template

bank is used. We show in the plot the ROC curve for a matched filter template bank with 95% average match, which is achieved with approximately 100000 templates. Since the matched filter is known to be statistically optimal for detecting a known signal in WGN, it is somewhat expected that this algorithm has the highest detection efficiency.

A potentially impactful difference between the matched filter and CNN algorithms is that the CNN relies upon convolutions as its fundamental calculation mechanism, whereas our implementation of a matched filter utilizes an inner product. Since convolution is a translation invariant operation, the detection performance of CNN can be extended to a wider range of CRES event kinetic energies with less cost than the matched filter, a feature that we exploited during the CNN training by including circular translations of the CRES frequency spectra in the training loop. Increasing the range of kinetic energies detectable by a matched filter requires a proportional increase in the number of templates, which directly translates into increased computational and hardware costs. From a practical perspective, the detection algorithm is always limited by the available computational hardware, so estimating the relative costs is a key factor in determining their feasibility. Below we perform a more detailed analysis of the relative costs of each of the detection algorithms.

#### 4.4.5.2 Computational Cost and Hardware Requirements

In the process of investigating triggering approaches for an antenna array CRES experiment, we have uncovered a strong tension between detection efficiency and computational resources. To relate the computational cost estimates to actual costs, we compare the theoretical amount of computer hardware required to implement the signal classifiers for real-time detection in an FSCD experiment. To do this we shall utilize order of magnitude estimates of the theoretical peak performance values for currently available Graphics Processing Units (GPUs) as a metric. This approach will underestimate the amount of required hardware, since it is unlikely that any CRES detection algorithm could reach the theoretical peak performance of the hardware.

Of the three detection algorithms tested, the power threshold classifier is the least expensive. It requires that we check whether the amplitude of each frequency bin in the STFT is below or above our decision threshold. The STFT combined with digital beamforming produces  $N_{\text{bin}}N_b$  frequency bins that must be checked every  $N_{\text{bin}}/f_s$  seconds. This requires approximately  $O(10^{10})$  FLOPS to check in real-time. Current generations of GPUs have peak theoretical performances in the range of  $O(10^{13}) - O(10^{14})$  FLOPS [46], dependent on the required floating-point precision of the computation. Therefore, the

entire computational needs of a real-time triggering system using a power threshold classifier, including digital beamforming and generation of the STFT, could be met by a single high-end GPU or a small number of less powerful GPUs. Since triggering is only one step of the full real-time signal reconstruction approach, limiting the computational cost of this stage is ideal. However, we have seen that the power threshold classifier does not provide sufficient detection efficiency across the entire range of possible signals, which is the primary motivation for exploring more complicated triggering solutions.

As discussed, the computational cost of the matched filter approach requires counting the number of templates that must be checked for each frequency spectra produced by the STFT. Computing the matched filter scores requires  $O(N_b N_t N_{\text{bin}})$  operations, since for each of the  $N_b$  beamforming positions we must multiply  $N_t$  templates with a data vector that has length  $N_{\text{bin}}$ . The time within which we must perform this calculation is equal to  $N_{\text{bin}}/f_s$  to keep up with the data generation rate. To cover the 5 eV kinetic energy range spanned by the template bank, we saw that  $10^4$  to  $10^5$  templates are required in order to match or exceed the detection efficiency of the CNN. If the number of templates scales linearly with the kinetic energy range of interest as expected, then we would require  $10^5$  to  $10^6$  matched filter templates with this more realistic range of energies. Considering this, the estimated computational cost of the matched filter is between  $O(10^{15})$  to  $O(10^{16})$  FLOPS, which is  $O(10^2)$  to  $O(10^3)$  high-end GPUs.

Lastly, we have the CNN classifier. To estimate the computational cost we simply sum the number of convolutions and matrix multiplications specified by the network architecture shown in Table 4.1. Each convolutional layer consists of  $N_{\text{in}} N_{\text{out}} N_{\text{kernel}} L_{\text{input}}$  floating-point operations, where  $N_{\text{in}}$  is the number of input channels,  $N_{\text{out}}$  is the number of output channels,  $N_{\text{kernel}}$  is the size of the convolutional kernel, and  $L_{\text{input}}$  is the length of the input vector, and the fully connected layers each contribute  $N_{\text{in}} N_{\text{out}}$  operations. Summing all the neural network layers we estimate that the CNN would require  $O(10^6)$  floating point operations for each frequency spectra; therefore, the total computation cost of the CNN trigger is this cost times the number of beamforming positions per the data acquisition time, which is  $O(10^{13})$  FLOPS or  $O(10^0)$  GPUs.

Compared with the matched filter approach the CNN requires  $O(100)$  to  $O(1000)$  fewer GPUs to implement, dependent on the exact number of templates used in the template bank. The 100 eV kinetic energy range is motivated by the application of these detection algorithms to an FSCD-like neutrino mass measurement experiment. However, if a significantly larger range of kinetic energies is required, a CNN may be the preferred detection approach despite the lower average detection efficiency due to computational

3024 cost considerations. The low estimated computational cost of the CNN is directly related  
3025 to the small network size.

3026 Additional experiments with larger CNNs, generated by increasing the depth and  
3027 width of the neural network, and we observed that these changes provided minimal  
3028 ( $\lesssim 1\%$ ) improvement in the classification accuracy of the model. A potential reason  
3029 for this could be the sparse nature of the signals in the frequency domain and the low  
3030 SNR which makes for a challenging dataset to learn from. Future work could investigate  
3031 modifications to the neural network architecture such as sparse convolutions, which may  
3032 improve the classification accuracy of the model or further reduce the computational  
3033 costs of this approach. Alternatively, more complicated CNN architectures such as a  
3034 ResNet [47] or VGG model [48] may provide improved classification performance over a  
3035 basic CNN. An additional promising area of investigation are recurrent neural networks,  
3036 which may be able to exploit the time-ordered features of the STFT for more accurate  
3037 signal detection if the electron signals last for multiple Fourier transform windows.

3038 Our estimate of the computational cost of the matched filter is somewhat naive if  
3039 we notice that the majority of the values that make up a CRES frequency spectra are  
3040 zero (see Figure 4.31). Therefore, the majority of operations in the matched filter inner  
3041 product are unnecessary, and we could instead evaluate the matched filter inner product  
3042 using only the  $\lesssim 10$  frequency peaks that make up CRES signal. This optimization  
3043 reduces the number of operations required to check each template by a factor of  $O(100)$   
3044 to  $O(1000)$ , which brings the estimated computational cost of the matched filter in  
3045 line with the CNN. Although this level of sparsity results in a multiplication with very  
3046 low arithmetic complexity, the resulting sparse matched filter algorithm is still likely  
3047 to be constrained by memory access speed rather than compute speed. Ultimately, the  
3048 comparison of the relative computational and hardware costs between the matched filter  
3049 and CNN will depend on the efficiency of the software implementation and hardware  
3050 support for neural network and sparse matrix calculations.

#### 3051 **4.4.6 Conclusion**

3052 Increasing the detection efficiency and overall event rate of the CRES technique represents  
3053 a key developmental path towards new scientific results and broader applications of the  
3054 CRES technique. It is what motivates both the antenna array detection approach and  
3055 the development of real-time signal reconstruction algorithms. We have demonstrated  
3056 that significant gains in the detection efficiency of the CRES technique are achievable  
3057 by utilizing triggering algorithms that account for the specific shape of CRES signals in

3058 the detector. These algorithms emphasize the need for accurate and fast methods for  
3059 CRES simulation, since they directly contribute to the success of matched filter methods  
3060 by providing a way to generate expected signal templates and also serve as a source of  
3061 training data for machine learning approaches.

3062 The improvements in detection efficiency offered by these alternative approaches to  
3063 triggering are crucial to the success of efforts to develop scalable technologies for CRES  
3064 measurement, since they provide a significant increase in the detectable parameter space  
3065 of CRES events, which allows for a better utilization of the larger detection volume.  
3066 While we have focused on the real-time detection of CRES signals from antenna arrays,  
3067 these same signal classifiers could be used in CRES experiments utilizing a different  
3068 detector technologies, since the same principles of signal detection will apply. For example,  
3069 previous CRES measurements by the Project 8 collaboration that utilized a waveguide  
3070 gas cell, could have improved their detection efficiency by employing a matched filter  
3071 or neural network classifier to identify trapped electrons with pitch angles that are too  
3072 small to be detected by the power threshold approach. Furthermore, alternative CRES  
3073 detector technologies such as resonant cavities [4] could also see similar improvements  
3074 in detection efficiency, which is of crucial importance to future efforts by the Project 8  
3075 collaboration to utilize CRES to measure the neutrino mass.

# **3076 Chapter 5**

# **3077 Antenna and Antenna Measurement Sys- 3078 tem Development for the Project 8 Ex- 3079 periment**

## **3080 5.1 Introduction**

**3081** The FSCD and antenna array CRES represent an innovative approach to beta-decay  
**3082** spectroscopy. While much can be learned from simulations about the systematics of  
**3083** CRES with antenna arrays, laboratory measurements and demonstrations provide critical  
**3084** inputs to sensitivity and simulation models as well as provide a means for calibration and  
**3085** commissioning of the experiment. Therefore, a robust program of antenna and antenna  
**3086** measurement hardware development is important to the success of the FSCD and the  
**3087** development of antenna array CRES more broadly.

**3088** In this chapter we summarize the development of an antenna measurement system  
**3089** at Penn State to implement and test the techniques of antenna array CRES on the  
**3090** bench-top, in order to support the efforts of the Project 8 collaboration. In Section 5.2  
**3091** we provide an introduction to some fundamental parameters and concepts related to  
**3092** antenna measurements as well as an overview of the Penn State antenna measurement  
**3093** system hardware. In Section 5.3 we include the manuscript of a paper [44] which  
**3094** details the design and characterization of a specialized antenna developed to mimic the  
**3095** electric fields emitted by an electron in a CRES experiment. This antenna, called the  
**3096** Synthetic Cyclotron Antenna (SYNCA), is intended as a calibration tool for antenna  
**3097** arrays developed for CRES measurements. Lastly, in Section 5.4 we summarize a set  
**3098** of prototype FSCD antenna array measurements with the SYNCA [7], which we use to  
**3099** validate the simulated performance of the antenna array and estimate systematic errors  
**3100** associated with the antenna array.

## 3101 5.2 Antenna Measurements for CRES experiments

### 3102 5.2.1 Antenna Parameters

3103 Antenna characterization measurements are intended to validate simulations of the  
3104 antenna array performance, which ultimately informs the neutrino mass sensitivity of  
3105 the experiment. In this section, I shall summarize a few fundamental concepts relating  
3106 to antennas and antenna measurement, before introducing how Project 8 uses antenna  
3107 measurement for the development of antenna array CRES.

#### 3108 5.2.1.1 Radiation Patterns

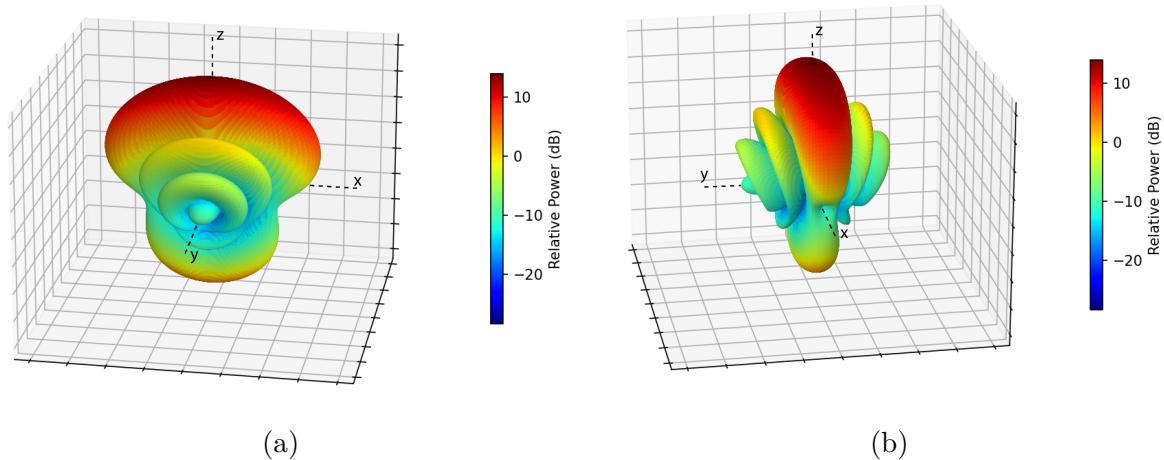


Figure 5.1: An example radiation pattern generated using HFSS simulations. The color and radial distance of the surface from the origin indicate the relative magnitude of radiation power emitted by the antenna in that direction. The primary goal of most antenna measurements is typically to measure the antenna pattern, which is used to derive many useful antenna performance parameters.

3109 Antennas are conductive structures designed to carry alternating electric currents  
3110 in order to transmit energy in the form of electro-magnetic (EM) waves [28]. Perhaps  
3111 the most fundamental way to characterize an antenna, is to map out the radiated power  
3112 density as a function of position, which is called the radiation pattern (see Figure 5.1).  
3113 We find the radiation power density by calculating the time-averaged Poynting vector for  
3114 all positions surrounding the antenna, which in equation form is

$$\mathbf{W}(x, y, z) = \langle \mathbf{E}(x, y, z, t) \times \mathbf{H}^*(x, y, z, t) \rangle_t, \quad (5.1)$$

3115 where  $\mathbf{E}(x, y, z, t)$  and  $\mathbf{H}(x, y, z, t)$  are the time-dependent electric and magnetic fields  
 3116 produced by the antenna [12]. The radiation power density has units of  $\text{W/m}^2$  and is  
 3117 more typically called the energy flux density in physics applications, since it is a measure  
 3118 of the amount of energy passing through a unit area over time.

3119 Because the radiation power density is a measure of power per unit area, its value  
 3120 in a particular direction will depend on the distance from the antenna at which we are  
 3121 measuring. This is undesirable for practical applications A related quantity, which is  
 3122 distance independent, is the energy flux per unit solid angle or radiation intensity, which  
 3123 is computed directly from the radition power density by multiplying by the squared  
 3124 distance from the antenna. Specifically,

$$U = r^2 W(x, y, z), \quad (5.2)$$

3125 where  $r$  is the distance from the antenna to the field measurement point. The radiation  
 3126 intensity is typically defined in regions where the Poynting vector consists only of a radial  
 3127 component where it is safe to treat as a scalar quantity.

### 3128 5.2.1.2 Directivity and Gain

3129 Since the radiation intensity is a measure of average power per unit solid angle, it is  
 3130 independent of distance and more useful as feature for antenna measurement. However,  
 3131 most antenna measurements are performed in terms of the directly related directivity  
 3132 and gain quantities. Directivity is defined as the ratio between the radiation intensity at  
 3133 particular point on the radiation pattern to the average radiation intensity computed  
 3134 over all solid angles [28]. The equation that relates the radiation intensity to directivity  
 3135 is

$$D = \frac{U}{U_0} = \frac{4\pi U}{P_{\text{rad}}}, \quad (5.3)$$

3136 where  $U_0$  is the average radiation intensity over all solid angles, which simply the total  
 3137 radiated power ( $P_{\text{rad}}$ ) divided by  $4\pi$ . Closely related to directivity is concept of gain,  
 3138 which accounts for energy losses that occur inside then antenna when attempting to  
 3139 transmit or receive a signal. The antenna gain is given by

$$G = \frac{4\pi U}{P_{\text{in}}}, \quad (5.4)$$

3140 where  $P_{\text{in}}$  is the total power delivered to the antenna. Gain can be thought of as the ratio  
 3141 of the antenna's radiation intensity to that of a hypothetical isotropic, lossless radiator.

<sup>3142</sup> The maximum values of gain and directivity exhibited by the main lobe of the antenna  
<sup>3143</sup> pattern as well as the ratio between the gain of the main lobe and any side-lobes are  
<sup>3144</sup> important figures of merit used to evaluate antenna designs.

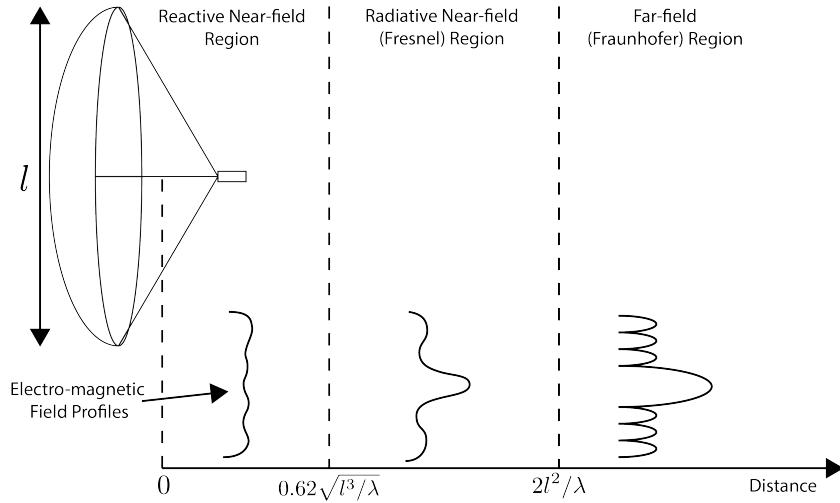


Figure 5.2: An illustration of the three field regions important for the analysis of an antenna system. Very close to the antenna the electric fields are primarily reactive so there is no radiation. If a receiving antenna were placed in this region most of the energy would be reflected back to the transmitter. Outside of the reactive near-field is the radiative near field. At these distances the antenna does radiate, but the radiation pattern is not well-defined since it changes based on the distance of the receiving antenna. It is only in the far-field region where the radiation pattern becomes constant as a function of distance, which is where the majority of antenna engineering is assumed to take place. The antenna arrays developed by Project 8 for CRES measurements operate in the radiative near-field due to the importance of limiting power loss from free-space propagation, which complicates the design of the antenna system.

### <sup>3145</sup> 5.2.1.3 Far-field and Near-field

<sup>3146</sup> Radiation patterns are only well-defined in regions where the shape of the radiation  
<sup>3147</sup> pattern is independent of distance. The region where this approximation is valid is called  
<sup>3148</sup> the "far-field", and in this region we can treat the EM fields from the antenna as spherical  
<sup>3149</sup> plane waves. A rule of thumb for antennas is that the far-field approximation is valid  
<sup>3150</sup> when the condition

$$R > \frac{2l^2}{\lambda} \quad (5.5)$$

<sup>3151</sup> is met. In this expression,  $R$  is the distance from the antenna,  $l$  is the largest characteristic  
<sup>3152</sup> dimension of the antenna, and  $\lambda$  is the wavelength of the radiation (see Figure 5.2).

3153        The region very close to the antenna is called the reactive near-field, because in this  
3154      region the reactive component of the EM field is dominant. Unlike radiative electric  
3155      fields, the reactive electric and magnetic fields are out of phase from each other by  
3156      90°, since they are the result of electrostatic and magnetostatic effects coming from the  
3157      self-capacitance and self-inductance of the antenna. The reactive fields are unable to  
3158      transfer energy a significant distance from the antenna and are thus completely negligible  
3159      for most antenna applications. The limit of the reactive near-field for an electrically-large  
3160      antenna is typically taken to be

$$R < 0.62\sqrt{l^3/\lambda}. \quad (5.6)$$

3161        The unique application of antennas by Project 8 is somewhat limited by reactive near-  
3162      field effects in the form of a maximum radial position for electrons inside the uniform  
3163      cylindrical antenna array. If electrons are too close to the edge of the array than reactive  
3164      near-field effects leads to a large reduction in the received power and consequently  
3165      detection efficiency. This leads to a significant volume inside of the antenna array that  
3166      is unsuitable for CRES lowering the volumetric efficiency of the antenna array CRES  
3167      technique relative to a cavity experiment.

3168        In between the reactive near-field and the far-field is the radiative near-field region.  
3169        In this region the fields are primarily radiative, however we are still too close to the  
3170      antenna for the spherical plane wave approximation to apply. Therefore, interference  
3171      effects between EM waves emitted from different points on the antenna occur causing the  
3172      shape of the radiation pattern to change as a function of distance from the antenna. If we  
3173      evaluate the far-field distance limit for the FSCD one finds an estimated far-field distance  
3174      of 43 cm, which is a factor of four larger than the radius of the antenna array designed for  
3175      the experiment. Consequently, we expect near-field effects to influence the performance  
3176      of the antenna array highlighting the importance of calibration and characterization  
3177      measurements.

#### 3178      **5.2.1.4    Polarization**

3179        The polarization of an EM wave defines the spatial orientation of the electric field  
3180      oscillations in the plane perpendicular to the direction of the propagation, and is defined  
3181      in terms of orthogonal polarization components. In our application, one analyzes the  
3182      properties of radiation propagating along the radial ( $\hat{r}$ ) direction away from the antenna,  
3183      which implies that the electric fields can be described as a linear combination of orthogonal

3184 polarization components

$$\mathbf{E}_{\text{tot}} = E_x \hat{x} + E_y \hat{y} + E_z \hat{z}, \quad (5.7)$$

3185 in Cartesian coordinates, or

$$\mathbf{E}_{\text{tot}} = E_\theta \hat{\theta} + E_\phi \hat{\phi}, \quad (5.8)$$

3186 in spherical coordinates.

3187 In general, one defines partial radiation patterns, directivities, and gains so that the  
3188 performance of the antenna for the desired polarization can be analyzed. The radiation  
3189 pattern defined in terms of partial patterns is

$$U_{\text{tot}} = U_\phi + U_\theta, \quad (5.9)$$

3190 where  $U_\phi$  and  $U_\theta$  are the radiation intensities in a particular direction for the respective  
3191 polarization components. Similarly, a quantity such as gain can be written in terms of  
3192 partial gains,

$$G_{\text{tot}} = G_\phi + G_\theta = \frac{2\pi U_\phi}{P_{\text{in}}} + \frac{2\pi U_\theta}{P_{\text{in}}}. \quad (5.10)$$

3193 If we view an electron performing a circular orbit in the XY-plane from the side, that  
3194 is, along the X or Y axes, then we would observe the electron to be performing a linear  
3195 oscillation perpendicular to the viewing axis. From this intuitive picture, we can predict  
3196 that the primary polarization of electric fields from CRES events to be linearly polarized  
3197 in the  $\hat{\phi}$  direction when viewed with an antenna positioned in the XY-plane.

### 3198 5.2.1.5 Antenna Factor and Effective Aperture

3199 A useful way to characterize the performance of an antenna is to measure the electric  
3200 field magnitude required to produce a signal with an amplitude of one volt in the antenna  
3201 terminals. This ratio between the magnitude of the incoming electric field and the  
3202 magnitude of the signal produced by the antenna is called the antenna factor, which is  
3203 written as

$$A_F = \frac{|\mathbf{E}_{\text{in}}|}{V_{\text{ant}}}, \quad (5.11)$$

3204 where  $A_F$  is the antenna factor,  $E_{\text{in}}$  is the incoming electric field, and  $V_{\text{ant}}$  is the magnitude  
3205 of the voltage produced by the antenna.

3206 The antenna factor can be expressed in terms of the antenna's gain through a related  
3207 quantity called the effective aperture. The effective aperture defines for a given incident  
3208 radiation power density ( $\text{W/m}^2$ ) the power that is received by the antenna. Therefore,

3209 the effective aperture gives the equivalent area of the antenna,

$$A_{\text{eff}} = \frac{P_{\text{rec}}}{P_{\text{in}}} = \frac{\lambda^2}{4\pi} G, \quad (5.12)$$

3210 where the received power is  $P_r$  and the total incoming power is  $P_{\text{in}}$ .

3211 If we express the incident power in terms of the magnitude of the Poynting vector,  
3212 then

$$|\mathbf{S}_{\text{in}}| = |\mathbf{E}_{\text{in}}|^2 / \eta_0, \quad (5.13)$$

3213 where  $\eta_0$  is the impedance of free-space, which relates the magnitudes of the electric and  
3214 magnetic fields in a vacuum, and is defined by

$$\eta_0 = \frac{|\mathbf{E}|}{|\mathbf{H}|} = \sqrt{\frac{\epsilon_0}{\mu_0}}. \quad (5.14)$$

3215 The total received power by the antenna can therefore be expressed as

$$P_{\text{rec}} = |\mathbf{S}_{\text{in}}| A_{\text{eff}} = |\mathbf{S}_{\text{in}}| \frac{\lambda^2}{4\pi} G = \frac{|\mathbf{E}_{\text{in}}|^2 \lambda^2 G}{4\pi \eta_0}. \quad (5.15)$$

3216 To relate this to the antenna factor recall that we can relate the voltage produced by  
3217 the antenna to the received power with

$$P_{\text{rec}} = \frac{V_{\text{ant}}^2}{Z} = \frac{|\mathbf{E}_{\text{in}}|^2}{A_F^2 Z}, \quad (5.16)$$

3218 where  $Z$  is the system impedance. Setting Equations 5.15 and 5.16 equal to each other,  
3219 we obtain the following expression for antenna factor in terms of gain

$$A_F = \sqrt{\frac{4\pi\eta_0}{ZG\lambda^2}} = \frac{9.73}{\lambda\sqrt{G}}. \quad (5.17)$$

3220 The second expression in Equation 5.17 is obtained by evaluating the constant terms  
3221 assuming a system impedance of  $50 \Omega$ .

3222 We have gone through the effort of expressing the antenna factor in terms of gain  
3223 to highlight that the majority of antenna parameters that we care to measure for a  
3224 CRES experiment can be obtained from the radiation or gain pattern of the antenna.  
3225 The antenna factor is a particularly important parameter for CRES measurements  
3226 due to its relevance to antenna array simulations with the Locust software [25, 43].  
3227 Specifically, Locust simulates the trajectory of an electron in a magnetic trap by running

3228 the Kassiopeia software package [23] and then uses the Liénard-Wiechert equations [26,27]  
3229 to calculate the electric fields that are incident on the antenna.

3230 To compute the response of the antenna to the electric field, Locust relies upon linear  
3231 time-invariant system theory [30], which computes the response of the antenna (i.e. the  
3232 voltage time series generated by the antenna) using a convolution between the electric field  
3233 time-series and the antenna impulse response. This approach is necessary for correctly  
3234 modeling the antenna response to the electric field due to the broadband and non-  
3235 stationary nature of the electric fields from CRES events. Since antenna measurements  
3236 take place under steady-state conditions, parameters such as the radiation pattern, gain,  
3237 and antenna factor are defined in the frequency domain. However, by performing an  
3238 inverse Fourier transform on the antenna factor we can obtain the antenna impulse  
3239 response, which allows us to simulate CRES events in the antenna array demonstrator  
3240 experiment.

### 3241 **5.2.2 Antenna Measurement Fundamentals**

#### 3242 **5.2.2.1 Friis Transmission Equation**

3243 The antenna factor, sometimes called the antenna transfer function, is used to model  
3244 how the antenna will respond to electric fields emitted from a CRES event. Therefore,  
3245 being able to measure the antenna transfer function of the antenna array is a key step  
3246 in the commissioning and calibration phases of an antenna array CRES experiment. A  
3247 common approach to antenna characterization is to perform a two antenna transmit-  
3248 receive measurement where an antenna with a known gain is used to characterize the  
unknown gain of the antenna under test (see Figure 5.3).

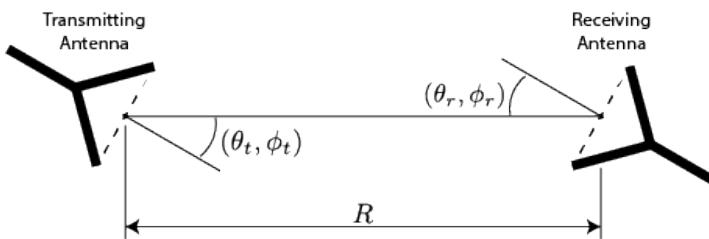


Figure 5.3: An illustration of the Friis measurement technique commonly used for antenna characterization measurements.

3249  
3250 To analyze this two antenna setup we seek to calculate the amount of power from  
3251 the transmitting antenna that we will detect with the receiving antenna. Using our  
3252 understanding of antenna gain, we can calculate the power density transmitted by an

3253 antenna in a direction  $(\theta_t, \phi_t)$  at frequency  $f$  and distance  $R$ , which is given by

$$w_t = \frac{P_t}{4\pi R^2} G_t(\theta_t, \phi_t, f). \quad (5.18)$$

3254 Here,  $P_t$  is the total power delivered to the transmitting antenna and  $G_t(\theta_t, \phi_t, f)$  is  
 3255 the value of the transmitting antenna gain. The power density is the power per unit  
 3256 area, so to calculate the total power delivered to the receiving antenna we multiply the  
 3257 transmitted power density by the effective area of the receiving antenna,

$$P_r = w_t A_{eff,r} = P_t \frac{G_t(\theta_t, \phi_t, f) G_r(\theta_r, \phi_r, f) c^2}{(4\pi R f)^2}, \quad (5.19)$$

3258 where  $G_r(\theta_r, \phi_r, f)$  is the gain of the receiving antenna. Equation 5.19 is called the  
 3259 Friis transmission equation [49, 50], which is of fundamental importance for antenna  
 3260 measurements, since it allows one to measure the gain of an unknown antenna by  
 3261 measuring the power received from an antenna with a known gain pattern. Alternatively,  
 3262 if no antenna with a known gain pattern is available, two identical antennas with unknown  
 3263 gain patterns can be used.

### 3264 5.2.2.2 S-Parameters and Network Analyzers

3265 Instead of directly measuring the power received by the antenna under test, it is more  
 3266 common to measure the ratio of the received power to the transmitted power,

$$\frac{P_r}{P_t} = \frac{G_t(\theta_t, \phi_t, f) G_r(\theta_r, \phi_r, f) c^2}{(4\pi R f)^2}. \quad (5.20)$$

3267 This power ratio can be easily measured using a vector network analyzer (VNA), which  
 3268 automates a significant fraction of the measurement process. Network analyzers are  
 3269 used to measure the scattering or S-parameters of a multi-port RF device [51], which  
 3270 describes how waves are scattered between the device ports. The antenna measurements  
 3271 we have been considering can be modeled as a two-port microwave device that we can  
 3272 characterize by measuring how incident voltage waves are transmitted or reflected (see  
 3273 Figure 5.4). We can write the scattered waves ( $V_1^-$  and  $V_2^-$ ) in terms of the incident ( $V_1^+$   
 3274 and  $V_2^+$ ) waves using the scattering matrix

$$\begin{pmatrix} V_1^- \\ V_2^- \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix} \begin{pmatrix} V_1^+ \\ V_2^+ \end{pmatrix}, \quad (5.21)$$

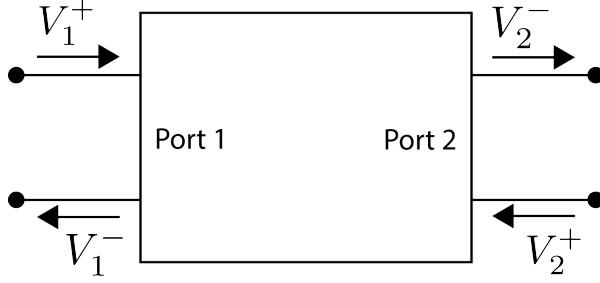


Figure 5.4: Illustration of a two-port S-parameter measurement setup. S-parameters characterize how incoming waves of voltage or power scatter off of the RF device under test. This allows you to measure important properties of the device. In particular, we can use this framework to model a two antenna radiation pattern measurement, which we can then automate using a VNA.

3275 where the elements of the matrix are the device S-parameters. It is assumed that,  
 3276 when exciting the device from a particular port, that all other ports in the network are  
 3277 terminated at the system impedance. This ensures that the incident waves from other  
 3278 ports in the network are zero. Therefore, the S-parameters are the ratios between the  
 3279 scattered and incident waves,

$$S_{ij} = \frac{V_i^-}{V_j^+}. \quad (5.22)$$

3280 Alternatively, S-parameters can be defined as the ratio of the scattered and incident  
 3281 power, which is proportional to the ratio of the squared voltage waves. Returning to  
 3282 our antenna measurement setup, we see that measuring the ratio of the received to the  
 3283 transmitted power is equivalent to measuring the ratio of power being scattered from port  
 3284 1 to port 2 in a RF network. Therefore, measuring an antenna's gain can be accomplished  
 3285 quite easily, by using a VNA to perform a two port  $S_{21}$  measurement.

### 3286 5.2.2.3 Antenna Array Commissioning and Calibration Measurements

3287 Up to this point we have been discussing calibration and commissioning measurements  
 3288 as they apply to a single antenna. While these measurements play an important role  
 3289 in validating the radiation patterns of the individual array elements, the ultimate goal  
 3290 is to use a phased array of these antennas. Therefore, we must also consider antenna  
 3291 measurement techniques that apply to the whole array system.

3292 By measuring the gain of each individual array element we can predict the features of  
 3293 the signals received during a CRES event using the antenna factor (see Section 5.2.1.5).  
 3294 However, unpredictable changes to the antenna performance can be introduced by the

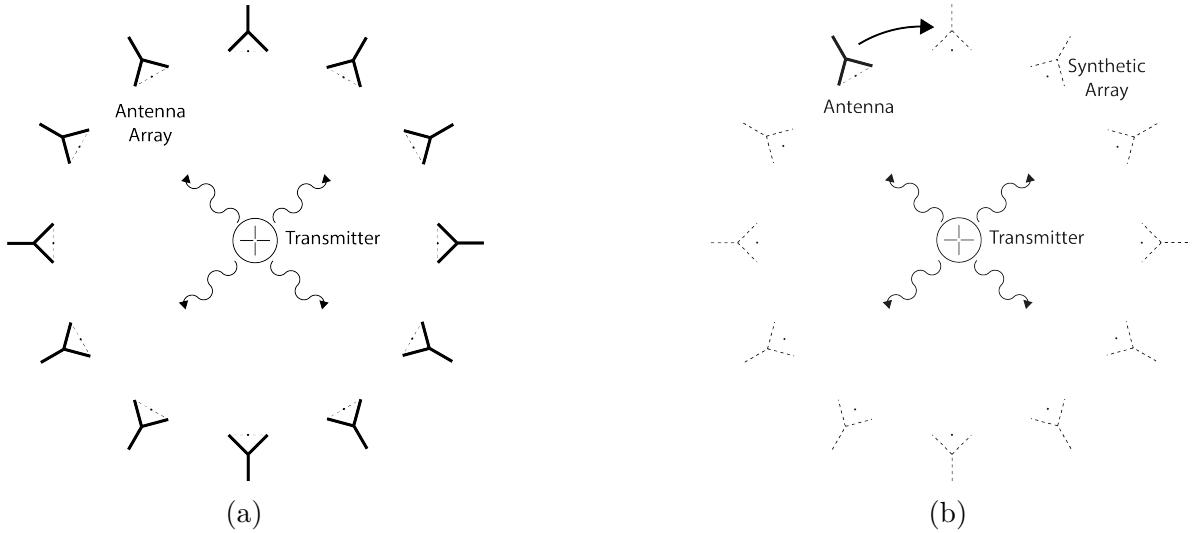


Figure 5.5: Two measurement approaches to characterizing an antenna array for CRES measurements. The full-array approach (a) requires a complete antenna array with all the associated hardware. The synthetic array approach (b) utilizes a single antenna and a set of rotation/translation stages to reposition the transmitter or the receiving antenna to synthesize the signals that would be received by the full-array. This approach reduces the cost and complexity of array measurements. A down-side of the synthetic array approach is that multi-channel effects such as reflections cannot be measured. Utilizing both the full-array and the synthetic array is a powerful way to quantify the impact of errors from the multi-channel array.

incorporation of the antennas into the circular array geometry, therefore, we employ both individual antenna and full-array measurements in the commissioning of the FSCD to account for these effects.

There are two main approaches to array measurements that could be used for characterization and calibration (see Figure 5.5). One approach is to construct the complete array and use an omni-directional transmitting antenna to measure the power received by each channel in the antenna array. In Section 5.3 we describe the development of an omni-directional transmitter that also mimics the radiation phase characteristics of a CRES event, which is useful because the entire array can be tested without repositioning. Alternatively, a full antenna array can be synthesized by repeatedly moving and measuring a single array element. This approach is ideal for identifying if different channels in the antenna array are affecting each other through multi-path interference by comparing the measurement results of the synthetic array to the real array.

### **5.2.3 The Penn State Antenna Measurement System**

The development of antenna array based CRES requires the capability to test and calibrate different antenna array designs to validate the performance of the as-built antenna array before and during the experiment. With these aims in mind we developed an antenna measurement system at Penn State specifically designed to mimic the characteristics of the antenna experiment designed for demonstration of the antenna array CRES technique by the Project 8 collaboration.

The Penn State antenna measurement system utilizes a two antenna measurement configuration with a stationary reference antenna and a test antenna mounted on a set of motorized translation and rotation stages (see Figure 5.6). The antenna measurement system can be operated in two distinct modes, one focused on the characterization of the radiation patterns of prototype antennas and the other focused on the validation of data-acquisition (DAQ) and CRES signal reconstruction techniques to bridge the gap between real measurements and simulation. In both measurement configurations it is critical to isolate the antennas from the environment so that multi-path reflections do not negatively influence the measurement results. For this reason we surround the measurement volume with microwave absorber foam (AEMI AEC-1.5) [52] specifically designed to attenuate microwave radiation near the 26 GHz measurement range of the system.

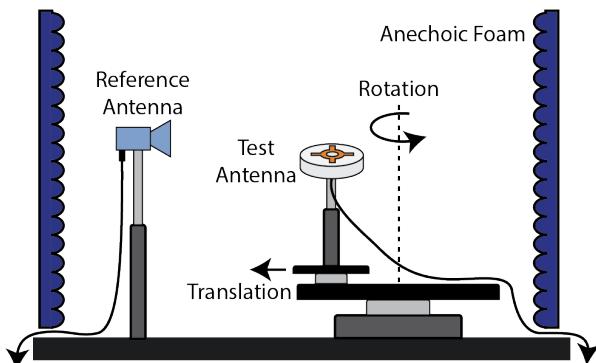


Figure 5.6: Illustration of the antenna measurement system developed for the Project 8 Collaboration. The reference and test antennas can be connected to different data acquisition configurations depending on the measurement goals. The reference antenna is typically a standard horn antenna and the test antenna is mounted on a set of translation stages for positioning. Automated translation stages allow for relatively painless data-taking enabling synthetic antenna array measurements using only a single receiving antenna. Anechoic form designed to mitigate RF reflections surrounds the setup.

In the first measurement configuration the reference antenna is typically a well-

3328 characterized horn antenna as pictured, since horn antennas have well-known and stable  
 3329 radiation patterns making them ideal as standard references. For characterization  
 3330 measurements, the test antenna represents the antenna-under-test whose pattern we wish  
 3331 to characterize. Mounting the test antenna on motorized rotation and translation stages  
 3332 allows us to automate the procedure significantly speeding up the radiation pattern  
 3333 measurement process.

3334 In the second measurement configuration one is interested in recreating the conditions  
 3335 of an antenna array CRES experiment as it concerns the antenna array and DAQ system.  
 3336 In this case, the reference antenna is a prototype FSCD antenna, which will be used to  
 3337 construct the antenna array in the FSCD experiment, and the test antenna is a specially  
 3338 designed synthetic cyclotron antenna (SYNCA) as picture in Figure 5.6. The SYNCA is  
 3339 designed such that the radiation pattern mimics that of a CRES electron so that the  
 3340 signals received by the prototype CRES array antenna mimic what is expected for a real  
 3341 CRES experiment.

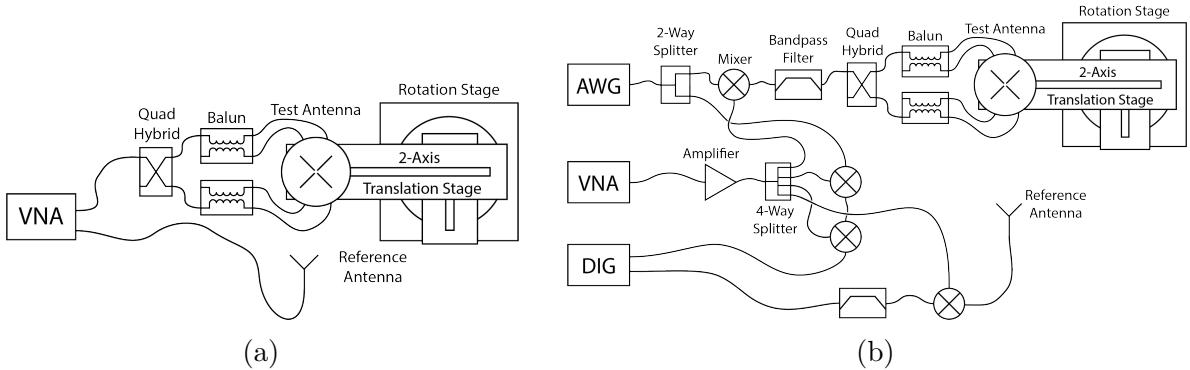


Figure 5.7: Diagrams of two measurement system configurations. Configuration (a) utilizes a VNA and is more suited to antenna characterization. Configuration (b) utilizes an AWG and VNA as a signal generation system and digitizer to collect measurement data, which is more suited to simulating CRES measurements. The transmission chain utilizes a quadrature hybrid and a pair of baluns to drive the cross-dipole variant test antenna developed for synthetic CRES measurements.

3342 In Figure 5.7 we show two high-level system diagrams of the Penn State antenna  
 3343 measurement system that depict the important system components and the connections  
 3344 between them. The two configurations of the measurement system utilize different  
 3345 hardware. For characterization and radiation pattern measurements, one prefers the  
 3346 configuration shown in Figure ???. In this case a vector network analyzer (VNA) is used  
 3347 as both the transmission source and data acquisition system and it is relatively easy to  
 3348 calibrate over a wide range of frequencies. Whereas, if one is more interested in recreating

3349 what would take place in the FSCD experiment then the configuration shown in Figure  
3350 ?? is preferable, since this system effectively mimics the receiver chain envisioned for the  
3351 FSCD experiment.

3352 The characterization configuration utilizes a network analyzer (Keysight N5222A)  
3353 [53, 54] with two independent sources and four measurement ports as the primary  
3354 measurement tool. A standard reference antenna is connected to one measurement port,  
3355 and the test antenna is connected to a separate port. The typical reference antenna  
3356 used for these studies is a Pasternack PF9851 horn antenna [55]. In the measurement  
3357 shown, the test antenna represents a SYNCA antenna, which requires a transmission  
3358 chain consisting of quadrature hybrid coupler [56, 57] (Marki QH-0226) connected to  
3359 two baluns [58] (Marki BAL-0026) to generate feed signals with the appropriate phases.  
3360 The VNA measures the radiation pattern by performing a transmission S-parameter  
3361 measurement, which can be used with the knowledge of the reference antenna's radiation  
3362 pattern to determine the radiation pattern of the test antenna (see Section 5.2.1).

3363 The second configuration is more complicated and incorporates more hardware  
3364 components in order to more closely mimic the DAQ system envisioned for the FSCD  
3365 experiment. The basic approach is to produce CRES-like radiation and use an antenna  
3366 combined with a realistic RF receiver chain to acquire the signals. On the transmit side,  
3367 an arbitrary waveform generator [59] (AWG, RIGOL DG5252) is used to generate a  
3368 waveform that mimics a CRES signal at a baseband frequency up to 250 MHz. This  
3369 frequency is then up-converted to the CRES signal frequency band of 25.8 to 26.0 GHz  
3370 using a mixer [60] (Marki MM1-0832L) and a bandpass filter (K&L Microwave 3C62-  
3371 25900/T200-K/K) to reject unwanted mixing components outside out of the 200 MHz  
3372 CRES signal band. The local oscillator signal for mixing is provided by one of the VNA  
3373 sources configured to run in a continuous wave setting. On the receive side, a prototype  
3374 antenna is used to detect the radiation emitted by the test antenna, which is down-  
3375 converted and filtered using the same mixer and bandpass filter as the transmission chain.  
3376 Lastly, data acquisition is performed using a 14-bit ADC sampling at 500 MSa/s [61]  
3377 (CAEN DT530) to digitize the down-converted signals.

3378 In order to distribute the LO to all mixers a 4-way power splitter (MiniCircuits  
3379 ZC4PD-18263-S+) along with an amplifier (Marki APM-6848) is used to drive the four  
3380 mixers used in the measurement system. A limitation of using the VNA as an LO source  
3381 is that there is no control of the LO phase when a measurement is triggered by the  
3382 control script, which leads to a random phase offset between acquisitions. This makes it  
3383 impossible to perform synthetic array measurements, which require strict control over

3384 the starting phase of the transmitted signal. In order to monitor the random phase of the  
3385 LO, a 2-way power splitter (MiniCircuits Z99SC-62-S+) is used to split the signal from  
3386 the AWG between the transmission path and a LO monitoring path. The LO monitoring  
3387 path consists of an up-conversion and down conversion using two mixers connected by a  
3388 coaxial cable, and monitors the relative phase of the LO using a channel on the digitizer  
3389 to sample this path. A phase shift in the LO will lead to a proportional phase shift in  
3390 the mixed signal, which is measured and removed from the received signals.

3391 The test antenna is mounted on a set of motorized stages, which are identical for  
3392 both measurement configurations. A rotational stage (ThorLabs PRMTZ8) is used as  
3393 the base layer with additional translation stages mounted on top of this. The rotational  
3394 stage is ideal for measuring a complete azimuthal scan of the test antenna's radiation  
3395 pattern as well as for moving a SYNCA antenna in circular motion to recreate the  
3396 symmetry of the FSCD antenna array. On top of the rotational stage we mount two  
3397 linear translation stages (ThorLabs MTS50-Z8 and MTS25-Z8) in a cross-wise manner  
3398 so that the test antenna can be moved along two perpendicular axes. Using the linear  
3399 stages in combination with the rotational stage allows one to fine-tune the positioning of  
3400 the test antenna so that it can be perfectly aligned with the central axis of the array.  
3401 A LabView script was developed to automate the measurement of a full 360° radiation  
3402 pattern and control the measurement electronics. Data from these acquisitions is stored  
3403 on university provided cloud storage.

## 3404 **5.3 Development of a Synthetic Cyclotron Antenna (SYNCA)** 3405 **for Antenna Array Calibration**

3406 This section is the manuscript of the publication [44] detailing the development of a  
3407 Synthetic Cyclotron Antenna (SYNCA) for antenna array characterization measurements  
3408 by the Project 8 collaboration.

### 3409 **5.3.1 Introduction**

3410 Neutrinos are the most abundant standard model fermions in our universe, but due to  
3411 weak interaction cross-sections with other particles, neutrinos are particularly difficult  
3412 to study. Consequently, many fundamental properties of neutrinos are still unknown  
3413 including the absolute scale of the neutrino mass [62]. Direct, kinematic measurements of  
3414 the neutrino mass are particularly valuable due to their model independent nature [63].

3415 To date the most sensitive direct neutrino mass measurements have been performed by  
 3416 the KATRIN collaboration [64], which measures the molecular tritium  $\beta$ -decay spectrum  
 3417 to infer the neutrino mass. Current data from neutrino oscillation measurements [62]  
 3418 allow for neutrino masses significantly smaller than the design sensitivity of the KATRIN  
 3419 experiment; therefore, there is a need for new technologies for performing direct neutrino  
 3420 mass measurements to probe lower neutrino masses.

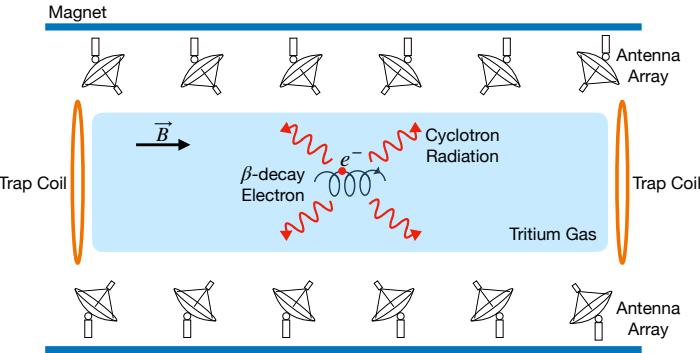


Figure 5.8: A sketch of an antenna array large-volume CRES experiment. Electrons from  $\beta$ -decays are confined in a magnetic field using a set of trap coils. The cyclotron radiation produced by the motion of the trapped electrons can be detected by a surrounding antenna array to determine the electron energies. Measuring the energies of many electrons produces a  $\beta$ -decay spectrum.

3421 The Project 8 collaboration is developing new methods for neutrino mass measurement  
 3422 based on Cyclotron Radiation Emission Spectroscopy (CRES) [19, 65–67], with the goal  
 3423 of measuring the absolute scale of the neutrino mass with a 40 meV/c<sup>2</sup> sensitivity [?, 63].  
 3424 This sensitivity goal will require the development of two separate technical capabilities.  
 3425 First is the development of an atomic tritium source, which avoids significant spectral  
 3426 broadening due to molecular final states [18]. Second is the technology for performing  
 3427 CRES in a multi-cubic-meter experimental volume with high combined detection and  
 3428 reconstruction efficiency, which is required in order to obtain sufficient event statistics  
 3429 near the tritium spectrum endpoint.

3430 One approach for a large-volume CRES experiment is to use an array of antennas,  
 3431 which surrounds a volume of tritium gas, to detect the cyclotron radiation produced  
 3432 by the  $\beta$ -decay electrons when they are trapped in a background magnetic field using a  
 3433 set of magnetic trapping coils (see Figure 5.8). Project 8 has developed a conceptual  
 3434 experiment design to study the feasibility of this approach. The design consists of a  
 3435 single circular array of antennas with a radius of 10 cm and 60 independent channels  
 3436 positioned around the center of the magnetic trap. The motivation behind this antenna

array design is to first develop an understanding of the antenna array approach to CRES  
 with a small scale experiment before attempting to scale the technique to large volumes  
 by using multiple antenna rings to construct the full cylindrical array. The development  
 of the antenna array approach to CRES has largely proceeded through simulations using  
 the Locust software package [43, 68], which is used to model the fields emitted by CRES  
 events and predict the signals received by the surrounding antenna array. To validate  
 these simulations, a dedicated test stand is being constructed to perform characterization  
 measurements of the prototype antenna array developed by Project 8 (see Figure 5.9)  
 and benchmark signal reconstruction methods using a specially designed transmitting  
 calibration probe antenna.

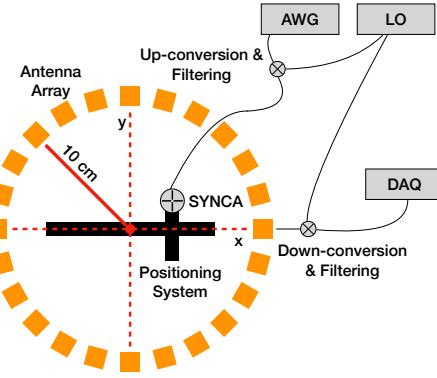


Figure 5.9: A schematic of the antenna array test stand. The circular antenna array has a radius of 10 cm with 60 independent channels (limited number shown for clarity). The test stand includes an arbitrary waveform generator (AWG), local oscillator (LO), and data acquisition (DAQ) hardware. Finally, a specialized Synthetic Cyclotron Antenna (SYNCA) is used to inject signals to test the antenna array.

We call this probe antenna the Synthetic Cyclotron Antenna or SYNCA. The SYNCA  
 is a novel antenna design that mimics the cyclotron radiation generated by individual  
 charged particles trapped in a magnetic field, which will be used in the antenna test  
 stand to perform characterization measurements, simulation validation, and reconstruc-  
 tion benchmarking. This paper provides an overview of the design, construction, and  
 characterization measurements of the SYNCA performed in preparation for its usage as  
 a transmitting calibration probe.

In Section 5.3.2 we provide a description of the cyclotron radiation field characteristics  
 that we recreate with the SYNCA. In Section 5.3.3 we give an overview of the simulations  
 performed to develop an antenna design that mimics the characteristics of cyclotron  
 radiation. In Section 5.3.4 we outline characterization measurements to validate that

3458 the fields generated by the SYNCA match simulation, and finally in Section 5.3.5 we  
3459 demonstrate an application of the SYNCA to test phased array reconstruction techniques  
3460 on the bench-top.

### 3461 5.3.2 Cyclotron Radiation Phenomenology

3462 To understand the cyclotron radiation phenomenology that the SYNCA should mimic,  
3463 we consider a charged particle moving at relativistic speed in the presence of an external  
3464 magnetic field (see Figure 5.10). In the special case we shall examine, the entirety of  
3465 the electron's momentum is directed perpendicular to the magnetic field; therefore, the  
3466 trajectory of the electron is confined to the cyclotron orbit plane. Because the momentum  
3467 vector is oriented perpendicular to the magnetic field, electrons with these trajectories  
3468 are said to have pitch angles of 90°.

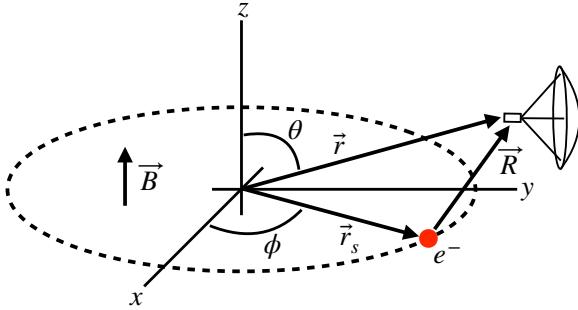


Figure 5.10: An electron (red dot) performing cyclotron motion in the x-y plane. The resulting cyclotron radiation is observed by an antenna located at the field point of interest.

3469 The cyclotron radiation fields generated by this circular trajectory are those which  
3470 we aim to reproduce with the SYNCA. We can describe the electromagnetic (EM) fields  
3471 using the Liénard-Wiechert equations [12, 43], which in non-covariant form express the  
3472 electric field as

$$\vec{E} = e \left[ \frac{\hat{n} - \vec{\beta}}{\gamma^2 (1 - \vec{\beta} \cdot \hat{n})^3 |\vec{R}|^2} \right]_{t_r} + \frac{e}{c} \left[ \frac{\hat{n} \times [(\hat{n} - \vec{\beta}) \times \dot{\vec{\beta}}]}{(1 - \vec{\beta} \cdot \hat{n})^3 |\vec{R}|} \right]_{t_r}, \quad (5.23)$$

3473 where  $e$  is the particle's charge,  $\hat{n} = (\vec{r} - \vec{r}_s)/|\vec{r} - \vec{r}_s|$  is the unit vector pointing from the  
3474 electron to the field measurement point,  $\vec{\beta} = \dot{\vec{r}}_s/c$  is the velocity of the particle divided  
3475 by the speed of light, and  $\gamma$  is the relativistic Lorentz factor. The equation is meant to  
3476 be evaluated at the retarded time as indicated by  $t_r = t - |\vec{R}|/c$ , which accounts for the

<sup>3477</sup> time delay due to the finite speed of light between the point where the field was emitted  
<sup>3478</sup> and the point where the field is detected.

<sup>3479</sup> We would like to simplify Equation 5.23 it at all possible. As a first step we analyze  
<sup>3480</sup> the relative magnitudes of the electric field polarization components. Consider an electron  
<sup>3481</sup> following a circular cyclotron orbit in a uniform magnetic field whose guiding center  
<sup>3482</sup> is positioned at the origin of the coordinate system. The equation of motion can be  
<sup>3483</sup> expressed as

$$\vec{r}_s = (r_c \cos \omega_c t_r) \hat{x} + (r_c \sin \omega_c t_r) \hat{y}. \quad (5.24)$$

<sup>3484</sup> For single antenna located along the y-axis at position  $\vec{r} = r_a \hat{y}$  we are interested in the  
<sup>3485</sup> incident electric fields from the electron. The electric field is given by Equation 5.23,  
<sup>3486</sup> which we evaluate in the regime where  $r_a \gg r_c$ . This limit can be justified by comparing  
<sup>3487</sup> the radius of the cyclotron orbit for an electron with the tritium beta-spectrum endpoint  
<sup>3488</sup> energy of 18.6 keV in a 1 T magnetic field to the typical ( $r_a \simeq 100$  mm) radial position  
<sup>3489</sup> of the receiving antenna. We find that the cyclotron orbit has a radius of 0.46 mm which  
<sup>3490</sup> is approximately a factor of 200 smaller than the typical antenna radial position. In this  
<sup>3491</sup> regime we can make the approximation  $\vec{R} \simeq r_a \hat{y}$  and the expression for the electric field  
<sup>3492</sup> at the antenna's position becomes

$$\vec{E} = \frac{e}{\gamma^2 r_a^2} \frac{\hat{x} \left( \frac{r_c \omega_c}{c} \sin \omega_c t_r \right) + \hat{y} \left( 1 - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)}{\left( 1 - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)^3} - \frac{e}{cr_a} \frac{\hat{x} \left( \frac{r_c^2 \omega_c^3}{c^2} - \frac{r_c \omega_c^2}{c} \cos \omega_c t_r \right)}{\left( 1 - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)^3}. \quad (5.25)$$

<sup>3493</sup> Since the receiving antenna is part of a circular array of antennas, it is useful to rewrite  
<sup>3494</sup> Equation 5.25 in terms of the azimuthal ( $\hat{\phi}$ ) and radial ( $\hat{r}$ ) polarizations. Making use of  
<sup>3495</sup> the fact that for an antenna located at  $R = r_a \hat{y}$  that  $\hat{\phi} = -\hat{x}$  and  $\hat{r} = \hat{y}$  we find

$$\vec{E} = \hat{\phi} E_\phi + \hat{r} E_r \quad (5.26)$$

$$E_\phi = \frac{e}{\left( 1 - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)^3} \left[ -\frac{\frac{r_c \omega_c}{c} \sin \omega_c t_r}{\gamma^2 r_a^2} + \frac{\omega_c \left( \frac{r_c^2 \omega_c^2}{c^2} - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)}{cr_a} \right] \quad (5.27)$$

$$E_r = \frac{e \left( 1 - \frac{r_c \omega_c}{c} \sin \omega_c t_r \right)}{\gamma^2 r_a^2 \left( 1 - \frac{r_c \omega_c}{c} \cos \omega_c t_r \right)^3}. \quad (5.28)$$

<sup>3496</sup> For the purposes of designing a synthetic cyclotron radiation antenna we are interested  
<sup>3497</sup> in the dominant electric field polarization emitted by the electron. The antenna is being  
<sup>3498</sup> designed to mimic the cyclotron radiation produced by electrons with kinetic energies of  
<sup>3499</sup> approximately 18.6 keV in a 1 T magnetic field [18]. Since the relativistic beta factor for

3500 an electron with this kinetic energy is  $|\vec{\beta}| \simeq \frac{1}{4}$ , the approximations  $\gamma \simeq 1$  and  $\frac{r_c \omega_c}{c} \simeq \frac{1}{4}$  are  
 3501 justified. Inserting these expressions into the equations for the electric field components  
 3502 above simplifies the comparison of the magnitudes of the two components. Additionally,  
 3503 we compare the time-averaged magnitudes to evaluate the root mean squared electric  
 3504 field ratio. The time-averaged ratio of the radial and azimuthally polarized electric fields  
 3505 with the above simplifications is given by

$$\frac{\langle |E_r| \rangle}{\langle |E_\phi| \rangle} = \frac{8 - \sqrt{2}}{\left| 1 - \frac{r_a}{r_c} \frac{1-2\sqrt{2}}{8} \right|} \simeq \frac{r_c}{r_a} \frac{8(8 - \sqrt{2})}{2\sqrt{2} - 1} = 0.13, \quad (5.29)$$

3506 where we have made use of the fact that for these magnetic fields and kinetic energies  
 3507 the cyclotron radius is much smaller than the radius of the antenna array.

3508 From Equation 5.29 we see that the time-averaged azimuthal polarization is larger than  
 3509 the radial polarization by about a factor of 8, which makes it the dominant contribution  
 3510 to the electric fields at the position of the antenna. We must also consider the directivity  
 3511 of the receiving antenna which can have a gain that is disproportionately large for a  
 3512 specific polarization component. Because the  $E_\phi$  component is dominant, the receiving  
 3513 antenna array is designed with an azimuthal polarization, which negates the voltages  
 3514 induced in the antenna from the radially polarized fields. Therefore, we conclude that  
 3515 for the purpose of designing the SYNCA antenna it is acceptable to approximate the  
 3516 electric fields from Equation 5.23 as purely azimuthally or  $\phi$ -polarized. The simplified  
 3517 expression for the electric field received by an antenna becomes

$$\vec{E} = E_\phi \hat{\phi} = \frac{e \frac{r_c \omega_c}{c}}{4r_a r_c} \left[ \frac{\frac{r_c \omega_c}{c} - \cos \omega_c t - \frac{4r_c}{r_a} \sin \omega_c t}{(1 - \frac{r_c \omega_c}{c} \cos \omega_c t)^3} \right]_{t_r} \hat{\phi}, \quad (5.30)$$

3518 where the radius of the cyclotron orbit is called  $r_c$ , the cyclotron frequency is called  $\omega_c$ ,  
 3519 and the radial position of the receiving antenna is called  $r_a$ . Equation 5.30 has been  
 3520 evaluated in the non-relativistic limit where  $\gamma \simeq 1$ , which is justified by the fact that  
 3521  $|\vec{\beta}| \simeq \frac{e}{4}$  for an electron with an 18.6 keV kinetic energy in a 1 T magnetic field.

3522 This rather complicated expression can be simplified using Fourier analysis. Assuming  
 3523 a background magnetic field of 1 T and a kinetic energy of 18.6 keV we calculate  
 3524 numerically the electric field using Equation 5.30 and apply a discrete Fourier Transform  
 3525 to visualize the frequency spectrum (see Figure 5.11).

3526 We observe that the azimuthally polarized electric field is periodic with a base cyclotron  
 3527 frequency of 25.898 GHz corresponding to the highest power frequency component in

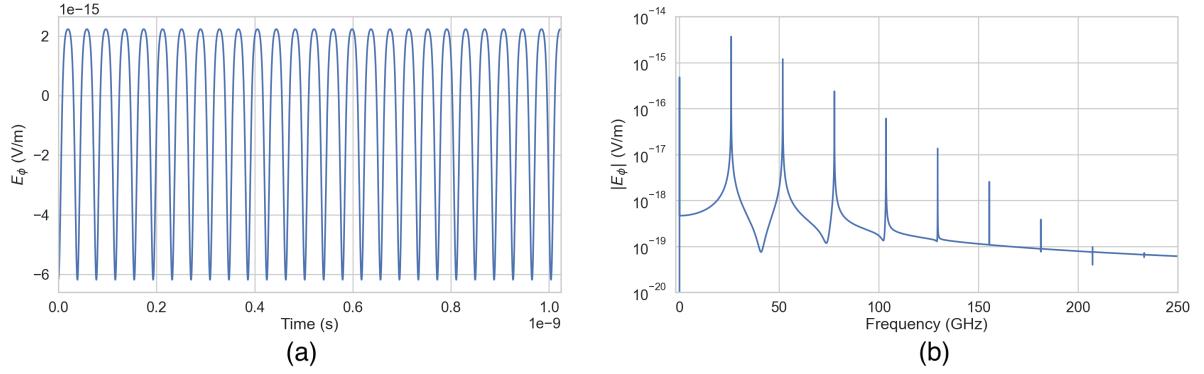


Figure 5.11: A plot of the numeric solution to Equation 5.31. The time-domain representation of the signal (a) is composed of a zero frequency term and a series of harmonics separated by the main cyclotron frequency as shown in the plot of the frequency spectrum (b). We can see that the relative amplitude of the harmonics beyond  $k = 7$  are smaller than the main carrier by a factor of about  $10^{-5}$  and are completely negligible.

3528 Figure 5.11. The frequency spectrum reveals that the signal is composed of a constant  
 3529 term with zero frequency and a series of harmonics separated by 25.898 GHz. Therefore,  
 3530 we can represent the azimuthal electric fields from the electron as a linear combination  
 3531 of pure sinusoids with frequencies given by  $\omega_k = k\omega_c$  ( $k \in 0, 1, 2, \dots$ ) and amplitudes  
 3532 extracted from the Fourier representation. Using this representation we can transform  
 3533 the equation for the azimuthally polarized electric fields in Equation 5.30 into

$$E_\phi = \frac{e^{\frac{r_c \omega_c}{c}}}{4r_a r_c} \sum_{k=0}^7 A_k e^{i\omega_k t_r}, \quad (5.31)$$

3534 where we have truncated the sum over harmonics at the 7th order for completeness. The  
 3535 amplitudes  $A_k$  are dimensionless complex numbers, which encode the relative powers of  
 3536 the harmonics as well as the starting overall phase of the cyclotron radiation. Because  
 3537 magnitude of the relative amplitudes exponentially decreases for higher harmonics, it is  
 3538 usually sufficient to consider only the terms up to  $k = 4$  where the relative amplitude  
 3539 of the harmonics has decreased from the main carrier by a factor of approximately 100.  
 3540 However, for completeness we include harmonics up to 7th order in Equation 5.31. The  
 3541 range of frequencies to which the receiving antenna array in the antenna test stand is  
 3542 sensitive is defined by the antenna's transfer function. The receptive bandwidth for  
 3543 the antennas used in the test stand is a range of frequencies with a bandwidth on the  
 3544 order of a few GHz centered around the main cyclotron carrier frequency of 25.898 GHz.  
 3545 Therefore, the higher order harmonics as well as the zero frequency term can be ignored

3546 when considering only the signals that will be received by the antenna array.

3547 Considering only the 1st order harmonic term from Equation 5.31, which represents  
3548 the portion of the electric field that will be detected by the array, and evaluating this at  
3549 the retarded time we obtain the following for the  $\phi$ -polarized electric fields

$$E_\phi \propto \cos \left( \omega_c \left( t - |\vec{R}|/c \right) - \Delta \right), \quad (5.32)$$

3550 where the arbitrary phase  $\Delta$  is defined by  $A_k = |A_k|e^{i\Delta}$ . We are interested in the  
3551 characteristics of the amplitude of the electric field as a function of the radial distance  
3552 component ( $|\vec{R}|$ ) of the retarded time. In particular, the maximum of  $E_\phi$  occurs when  
3553 the argument of the cosine function is equal  $n\pi$  where  $n \in \{0, \pm 2, \pm 4, \dots\}$ ; however, the  
3554 solutions where  $n$  is negative can be discarded since they represent unphysical negative  
3555 overall phases. Applying this condition to Equation 5.32 gives a condition on the radial  
3556 position of the maximum of  $E_\phi$

$$\omega_c(t - |\vec{R}|/c) - \Delta = n\pi, \quad (5.33a)$$

$$|\vec{R}| = \frac{c}{\omega_c} ((\omega_c t - \Delta) - n\pi), \quad (5.33b)$$

3557 which is a function of time in the frame of the moving electron ( $t$ ). Equation 5.33 can  
3558 be further simplified by noticing that the azimuthal position of the electron ( $\phi_e(t)$ ) as a  
3559 function of time is defined by  $\phi_e(t) = \omega_c t - \Delta$  which reduces Equation 5.33 to

$$|\vec{R}| = \frac{c}{\omega_c} (\phi_e(t) - n\pi). \quad (5.34)$$

3560 Equation 5.34 represents an archimedean spiral which is formed when plotting the  
3561 amplitude of  $E_\phi$  in the x-y plane. The solution where  $n = 0$  represents the leading edge  
3562 of the radiation spiral which propagates outward from the electron at the speed of light.  
3563 The additional solutions for  $n > 0$  represent the persistent spiral at radii inside the  
3564 leading edge of the radiated fields that have not yet been detected by the receiver at the  
3565 current time. In Figure 5.12a we show the expected spiral pattern for the maxima of the  
3566 cyclotron radiation.

3567 In particular, we note that for the circular array geometry of the test stand, depicted  
3568 as the series of circles in Figure 5.12a, each antenna receives a linearly polarized wave  
3569 with a phase offset that corresponds to the azimuthal angle for that antenna element.  
3570 Therefore, as we show in Figure 5.12b, when the relative phase of the received signal is  
3571 plotted as a function of the receiving antenna's azimuthal position the result is also an

3572 Archimedean spiral.

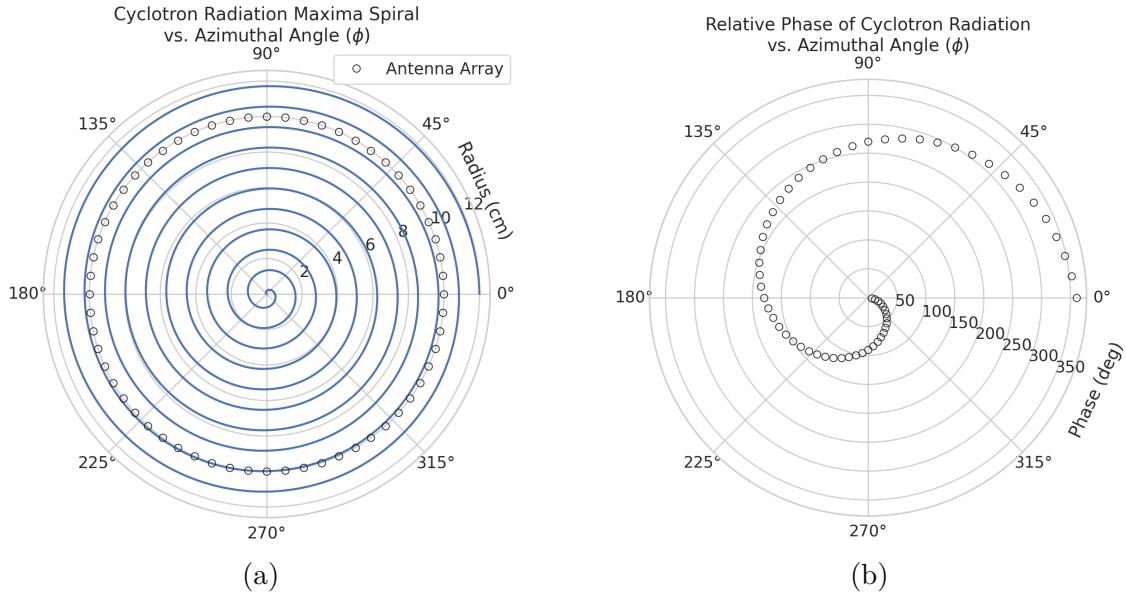


Figure 5.12: The amplitude maxima of the cyclotron radiation form an Archimedean spiral as the radiation propagates outward from the cyclotron orbit center (a). A circular antenna array located at a fixed radius from the orbit center will receive electric fields with equal magnitude in each of its channels, but the phase of the electric field incident on each array channel will be linearly out of phase from its neighbor antennas by an amount equal to the angular separation of the two channels (b).

3573 Based on these analytical calculations we can characterize the magnitude, polarization,  
3574 and phase of the signals received by the antenna array using three criteria. These criteria  
3575 are the basis of comparison for the radiation produced by the SYNCA and cyclotron  
3576 radiation emitted by electrons and will be used to evaluate the performance of antenna  
3577 designs. The criteria are:

- 3578 1. Electric fields that are  $\phi$ -polarized near  $\theta = 90^\circ$
- 3579 2. Uniform time-averaged electric field magnitudes around the circumference of a  
3580 circle centered on the antenna
- 3581 3. Electric fields whose phase is equal to the azimuthal angle at the point of measure-  
3582 ment plus a constant

3583 The Locust simulation package [68] can be used to directly simulate the EM fields  
3584 generated by electrons performing cyclotron motion to validate the analytical calculations.  
3585 Locust simulates the EM fields by first calculating the trajectory of the electrons in

3586 the magnetic trap using the Kassiopeia software package [69]. The trajectory can then  
 3587 be used to solve for the EM fields using the Liénard-Wiechert equations directly with  
 3588 no approximations. The resulting electric field solutions drive a receiving antenna by  
 3589 convolving the time-domain fields with the finite-impulse response filter of the antenna  
 3590 or they can be examined directly to study the field characteristics that the SYNCA must  
 3591 reproduce. In the next section we compare the radiation field patterns for electrons  
 3592 simulated with Locust to patterns from a SYNCA antenna design.

### 3593 5.3.3 SYNCA Simulations and Design

3594 One potential SYNCA design is the crossed-dipole antenna [70]. A crossed-dipole antenna  
 3595 consists of two dipole antennas, one of which is rotated 90° with respect to the other,  
 3596 which are fed with signals that are out of phase from the opposite dipole by 90° (see  
 Figure 5.13). This arrangement causes the signals fed to each arm of the dipole to be

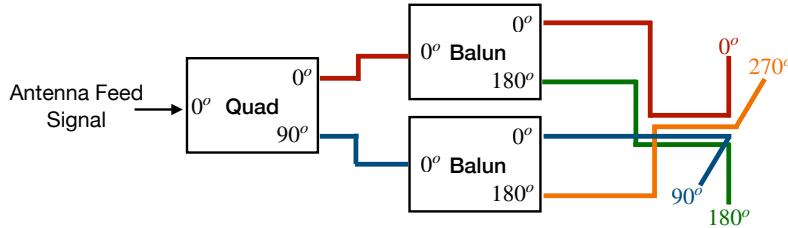


Figure 5.13: An idealized crossed-dipole antenna consists of two electric dipole antennas oriented perpendicular to each other and is fed with four signals with a quadrature phase relationship. An example antenna feed circuit is shown which is composed of a chained combination of a quadrature hybrid-coupler (Quad) and two baluns.

3597  
 3598 out of phase from each of the neighboring arms by 90°, which mirrors the spatial phase  
 3599 relationship of cyclotron radiation fields.

3600 A potential drawback of this design is that standard crossed-dipole antennas do not  
 3601 radiate uniform electric fields near the  $\theta = \pi/2$  plane. Typical crossed-dipole antennas  
 3602 use dipole arm lengths equal to  $\lambda/4$  or larger [70], where  $\lambda$  is the wavelength at the  
 3603 desired operating frequency. Such large arm lengths cause the electric field magnitude  
 3604 to vary significantly around the circumference of the antenna. However, making the  
 3605 antenna electrically small by shrinking the arm length can improve the antenna pattern  
 3606 uniformity.

3607 In general, the criterion for an electrically small antenna is that the largest dimension  
3608 of the antenna ( $D$ ) obey  $D \lesssim \lambda/10$  [28]. In our application, we are attempting to mimic  
3609 the cyclotron radiation emitted by electrons produced from tritium  $\beta$ -decay with energies  
3610 near the spectrum endpoint. For a background magnetic field of 1 T, the corresponding  
3611 cyclotron frequency of tritium endpoint electrons is approximately 26 GHz. Therefore, the  
3612 electrically small condition would require that the largest dimension of the crossed-dipole  
3613 antenna be smaller than 1.2 mm.

3614 A crossed-dipole antenna with an overall size of 1.2 mm is challenging to fabricate due  
3615 to the small dimensions of the dipole arms that, in practice, are fragile and unsuitable  
3616 for use as a calibration probe. To mitigate some of the challenges with the fabrication  
3617 of such a small antenna, a variant crossed-dipole antenna design using printed circuit  
3618 board (PCB) technology (see Figure 5.14) was developed in partnership with an antenna  
prototyping company, Field Theory Consulting <sup>1</sup>.

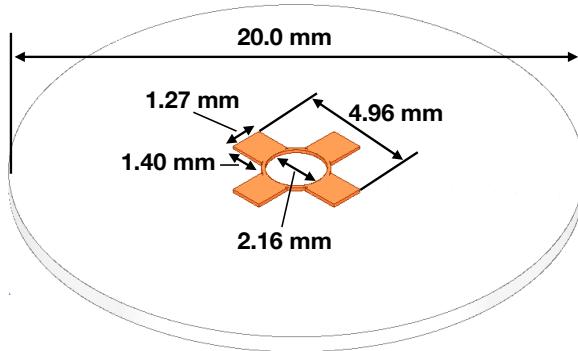


Figure 5.14: A model of the PCB crossed-dipole antenna with dimensions. The design has an inside diameter of 2.16 mm for the central circular trace, which is 0.13 mm wide. The dipole arms each have a width of 1.27 mm and protrude beyond the circular trace by 1.40 mm, which gives an overall width of 4.96 mm for the length of the antenna PCB trace from end-to-end. The overall size of the antenna is 20.0 mm the majority of which is the PCB dielectric material. This design was observed in simulation to maintain the field characteristics of the idealized crossed-dipole while being simpler to fabricate due to the increased size of the antenna.

3619  
3620 The PCB crossed-dipole design uses four rectangular pads to represent the dipole arms,  
3621 which are connected by a thin circular trace. The circular trace both adds mechanical  
3622 stability to the antenna and improves the azimuthal uniformity of the electric fields  
3623 compared to a more standard crossed-dipole geometry. Furthermore, the circular trace  
3624 allows for a greater separation between dipole arms than standard crossed-dipoles, which

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<sup>1</sup><https://fieldtheoryinc.com/>

3625 is required to accommodate the coaxial connections to each pad. The pads each contain  
 3626 a through-hole solder joint to connect coaxial transmission lines using hand soldering.  
 3627 The antenna PCB has no ground plane on the bottom layer as this was observed in  
 3628 simulation to significantly distort the radiation pattern in the plane of the PCB. The  
 3629 only ground planes present in the model are the outer conductors of the four coaxial  
 3630 transmission lines which feed the antenna. These are left unterminated on the bottom of  
 3631 the PCB dielectric material.

3632 The antenna design development utilized a combination of Locust electron simula-  
 3633 tions and antenna simulations using ANSYS HFSS [29], a commercial finite-element  
 3634 electromagnetic simulation software. Two antenna designs were simulated: an idealized  
 3635 electrically small crossed-dipole antenna with an arm length of 0.40 mm and an arm  
 3636 separation of 0.05 mm, as well as a PCB crossed-dipole antenna with the dimensions  
 3637 shown in Figure 5.14. Plotting the magnitude of the electric fields generated by the  
 3638 antennas across a 10 cm square located in the same plane as the respective antennas  
 3639 reveals the expected cyclotron spiral pattern (see Figure 5.15) which closely matches  
 3640 the prediction for simulated electrons. The spiral pattern demonstrates that the electric  
 3641 fields have the appropriate phases to mimic cyclotron radiation, which fulfills SYNCA  
 criterion 3 identified in Section 5.3.2.

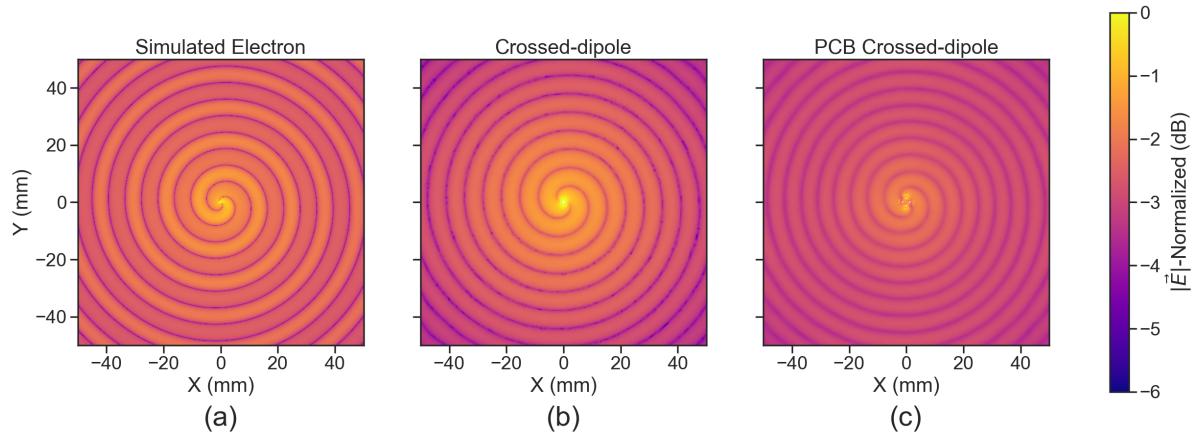


Figure 5.15: A comparison of the electric field magnitudes, normalized by the maximum value of the electric field in each simulation, plotted on a 10 cm square to visualize the Archimedean spirals formed by the electron (a), the crossed-dipole antenna (b), and a PCB crossed-dipole antenna (c). The matching patterns indicate that the electric fields have similar phase characteristics. These images were generated using Locust simulations for the electron and ANSYS HFSS for both antennas.

3642

3643 As we can see from Figure 5.16, the crossed-dipole antenna, which uses an idealized

3644 geometry, exhibits good agreement with simulation. The antenna has a maximum  
 3645 deviation from a simulated electron of approximately 0.5 dB in the total electric field, 1  
 3646 dB for the  $\phi$ -polarized electric field and 1 dB for the  $\theta$ -polarized electric field.

3647 In comparison, the pattern of the PCB crossed-dipole antenna, because the simulation  
 3648 incorporates the geometry of the coax transmission lines, exhibits some distortion from  
 3649 the idealized cross-dipole simulations. The vertically oriented ground planes of the coax  
 3650 lines introduce more  $\theta$ -polarized electric fields than are observed for simulated electrons  
 3651 near  $\theta = 90^\circ$ . The significant  $\theta$ -polarized field minimum is still present but shifted  
 to approximately  $\theta = 65^\circ$ . The  $\theta$ -polarized field deviations of the PCB crossed-dipole

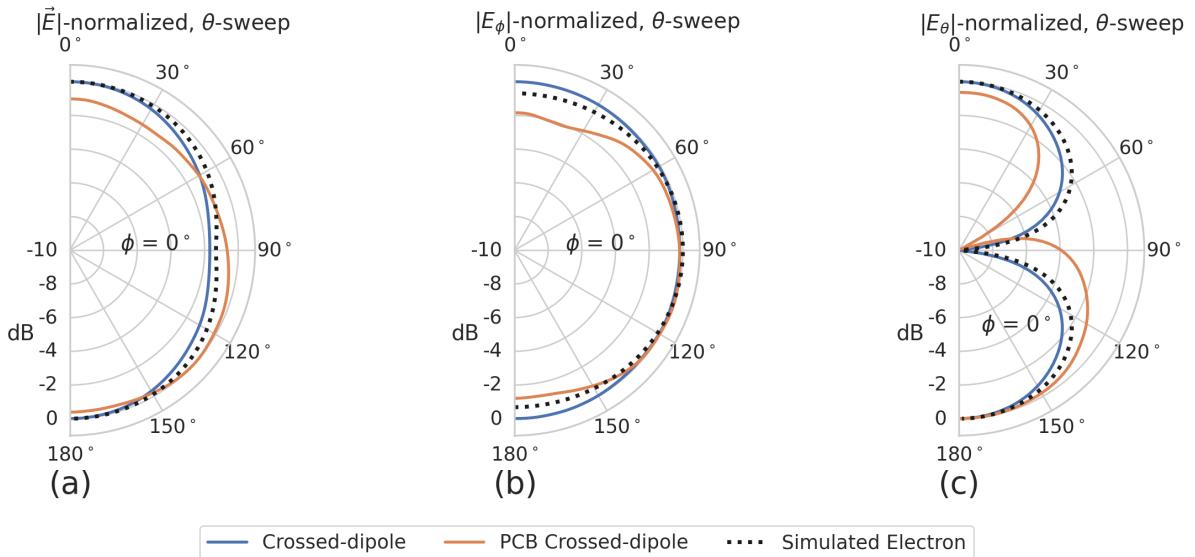


Figure 5.16: A comparison of the normalized electric field magnitudes for the ideal crossed-dipole, PCB crossed-dipole, and a simulated electron as a function of the polar angle ( $\theta$ ). (a) Shows the total electric field, (b) shows the  $\phi$ -polarized electric field component, and (c) shows the  $\theta$ -polarized electric field component. These images were generated using Locust simulations for the electron and ANSYS HFSS for both antennas.

3652  
 3653 antenna should not greatly impact the performance of the antenna because the receiving  
 3654 antenna array is primarily  $\phi$ -polarized. Therefore deviations in the  $\theta$ -polarized fields  
 3655 will be suppressed due to the polarization mismatch. More importantly, the  $\phi$ -polarized  
 3656 electric field pattern generated by the PCB crossed-dipole closely matches simulated  
 3657 electrons across the polar angle range of  $50^\circ < \theta < 150^\circ$ . In this region the PCB crossed-  
 3658 dipole differs by less than 0.5 dB from simulated electrons. This range greatly exceeds  
 3659 the beamwidth of the receiving antenna array which is designed to be most sensitive  
 3660 to fields produced near  $\theta = 90^\circ$ . Therefore, we conclude that the PCB crossed-dipole

3661 antenna generates a  $\phi$ -polarized radiation pattern that fulfills SYNCA criterion 1 from  
3662 Section 5.3.2.

3663 The final SYNCA criterion is related to the uniformity of the electric fields when  
3664 measured azimuthally around the antenna. As we saw for real electrons in Section 5.3.2  
3665 it is expected that the magnitude of the electric field be completely uniform as a function  
3666 of the azimuthal angle due to the symmetry of the cyclotron orbit. In Figure 5.17 we plot  
3667 the total electric field as a function of azimuthal angle for an electron, the crossed-dipole  
antenna, and the PCB crossed-dipole antenna. The crossed-dipole antenna exhibits

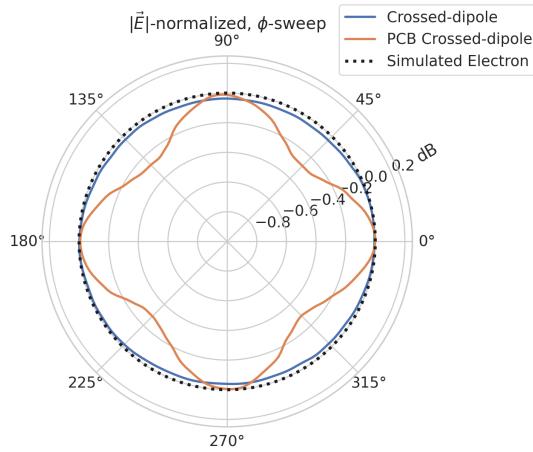


Figure 5.17: A comparison of the normalized electric field magnitudes for the crossed-dipole, PCB crossed-dipole, and a simulated electron as a function of the azimuthal angle ( $\phi$ ) evaluated at  $\theta = 90^\circ$ . This image was generated using Locust simulations for the electron and ANSYS HFSS for both antennas.

3668  
3669 perfect uniformity around the azimuthal angle, whereas the PCB crossed-dipole has a  
3670 small periodic deviation with a maximum difference of 0.3 dB caused by the coaxial  
3671 transmission lines below the PCB. Such a small deviation from uniformity is acceptable  
3672 since it is smaller than the expected variation in uniformity caused by imperfections in  
3673 the antenna fabrication process, which modifies the antenna shape in an uncontrolled  
3674 manner by introducing solder blobs with a typical size of a few tenths of a millimeter on  
3675 the dipole arms (see Figure 5.18). Additionally, the SYNCA will be separately calibrated  
3676 to account for azimuthal differences in the electric field magnitude. Therefore we see  
3677 from the simulated performance of the PCB crossed-dipole antenna that this antenna  
3678 design meets all three of the SYNCA criteria.

### 5.3.4 Characterization of the SYNCA

Two SYNCAs were manufactured using the PCB crossed-dipole design (see Figure 5.18). The antenna PCB (Matrix Circuit Board Materials, MEGTRON 6) is connected to four 2.92 mm coaxial connectors (Fairview Microwave, SC5843) using semi-rigid coax (Fairview Microwave, FMBC002), which also physically support the antenna PCB. The antenna PCB consists only of two layers which correspond to the copper antenna trace and the PCB dielectric. Each coax line is connected to the associated dipole arm using through-hole soldering and phase matched to ensure that the electrical length of each of the transmission lines is identical at the operating frequency. The antenna PCB is further reinforced using custom cut polystyrene foam blocks, which have an electrical permittivity nearly identical to air. A custom 3D printed mount is included at the base of the antenna to support the coax connectors and to provide a sturdy mounting base.

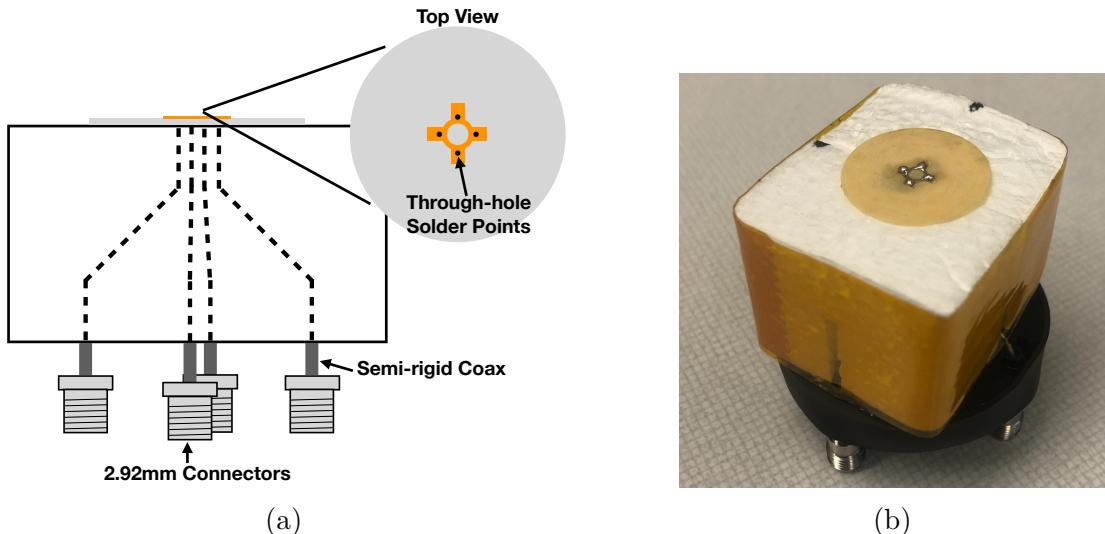


Figure 5.18: (a) A cartoon schematic which highlights the routing of the semi-rigid coax transmission lines. (b) A photograph of a SYNCA constructed using the modified crossed-dipole PCB antenna design. Visible in the photograph of the SYNCA are four blobs of solder which are an artifact of the SYNCA's hand-soldered construction. These solder blobs are the most significant deviation from the SYNCA design shown in Figure 5.14 and are responsible for a significant fraction of the irregularities seen in the antenna pattern.

Characterization measurements were performed using a Vector Network Analyzer (VNA) to measure the electric field magnitude and phase radiated by the SYNCA to verify the radiation pattern (see Figure 5.19). The VNA is connected to the SYNCA

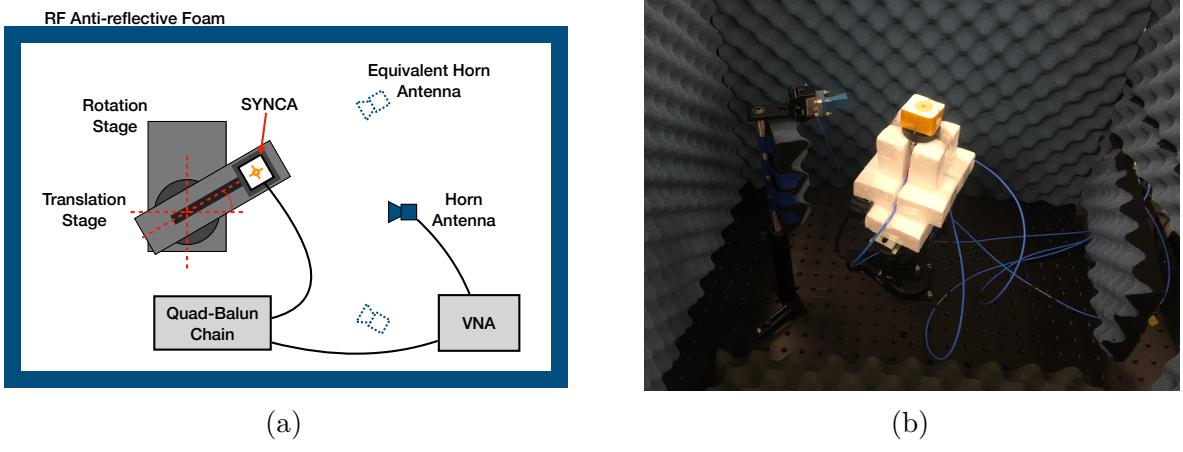


Figure 5.19: A schematic of the VNA characterization measurements (a). This setup allows for antenna gain and phase measurements across a full  $360^\circ$  of azimuthal angles using a motorized rotation stage and control of the radial position of the SYNCA using a translation stage. A photo of the setup in the lab is shown in (b).

3694 at one port through a hybrid-coupler whose outputs are connected to two baluns to  
 3695 generate the signals with the appropriate phases to feed the SYNCA (see Figure 5.13).  
 3696 The other port of the VNA is connected to a single reference horn antenna that serves  
 3697 as a field probe. To position the SYNCA, a combination of translation and rotation  
 3698 stages are used to characterize the antenna's fields across the entire radiation pattern  
 3699 circumference. This measurement scheme is equivalent to measuring the fields generated  
 3700 by the SYNCA using a full circular array of probe antennas.

3701 The antenna measurement space is surrounded by RF anti-reflective foam to isolate  
 3702 the measurements from the lab environment (see Figure 5.19b) and remaining reflections  
 3703 are removed using the VNA's time-gating feature. The SYNCA is affixed to the stages  
 3704 by a custom RF transparent mount made of polystyrene foam. The coaxial cables deliver  
 3705 the antenna feed signals generated by the VNA to the SYNCA while still allowing  
 3706 unrestricted rotation. The horn antenna probe is nominally positioned in the plane  
 3707 formed by the antenna PCB ( $\theta = 90^\circ$  or  $z = 0$  mm) at a distance of 10 cm from the  
 3708 SYNCA, to match the expected position of the antenna array relative to the SYNCA in  
 3709 the antenna array test stand. The horn antenna can be manually raised or lowered to  
 3710 different relative vertical positions to characterize the radiation pattern at different polar  
 3711 angles.

3712 Several  $360^\circ$  scans were performed with probe vertical offsets of -10.0 mm, -5.0 mm,  
 3713 0.0 mm, 5.0 mm, and 10.0 mm relative to the antenna PCB plane. These probe offsets

3714 cover a 2 cm wide vertical region centered on the SYNCA PCB, approximately equal to  
 3715  $\pm 6$  degrees of polar angle. The measurements show that the SYNCA is generating fields  
 3716 with nearly isotropic magnitude across the probed region. The standard deviation of the  
 3717 electric field magnitude measured around the antenna circumference is approximately  
 3718 2.9 dB for a typical rotational scan. The presence of a significant pattern null is noted  
 3719 near 45° (see Figure 5.20), which we attribute to small imperfections in the antenna  
 3720 PCB that could be introduced from the hand soldered terminations connecting the coax  
 3721 cables to the antenna. There is no significant difference in the radiation pattern when  
 3722 measured across the 2 cm vertical range. The measured relative phases closely follow  
 3723 the expectation for an electron, being linear with the measurement rotation angle and  
 3724 forming the expected spiral pattern. Other than the small phase imperfections there is  
 3725 a slight sinusoidal bias to the phase data, which we determined is the result of a small  
 3726 ( $\lesssim 1$  mm) offset of the antenna's phase center from the rotation axis of the automated  
 3727 stages.

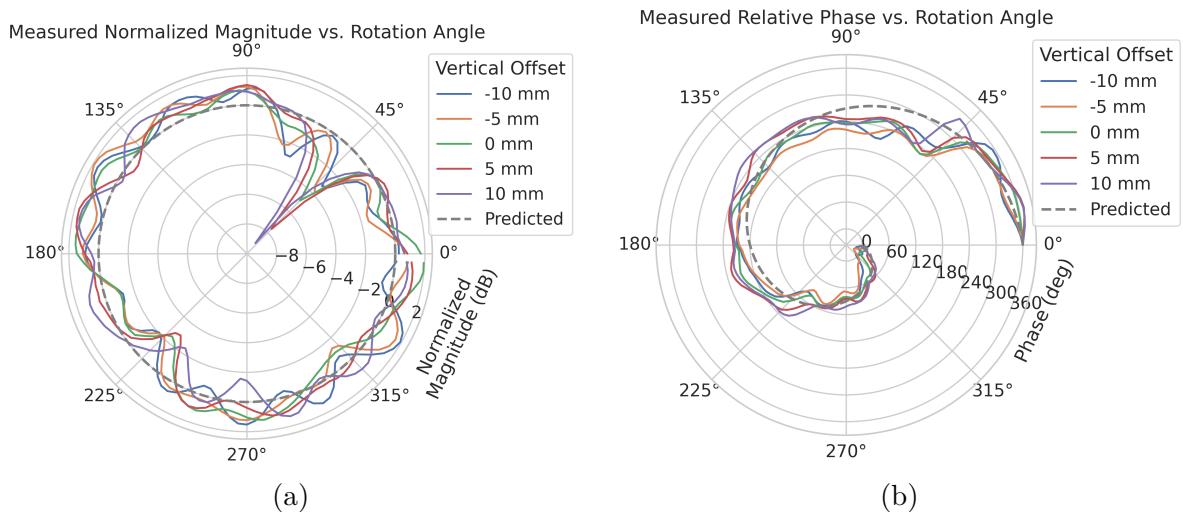


Figure 5.20: Linear interpolations of the measured electric field magnitude (a) and phase (b). The data was acquired using a VNA at 120 points spaced by 3 degrees from 0 to 357 degrees of azimuthal angle. The different color lines indicate the vertical offset of the horn antenna relative to the SYNCA PCB and the dashed line shows the expected shape from electron simulations. No significant difference in the antenna pattern is observed for the measured vertical offsets.

3728 The characterization measurements confirm the simulated performance of the SYNCA.  
 3729 As expected the fields generated by the antenna are nearly isotropic in magnitude,  $\phi$ -  
 3730 polarized, and are linearly out of phase around the circumference of the antenna as

3731 predicted for cyclotron radiation in Section 5.3.2. Small imperfections in the magnitude  
 3732 and phase of the antenna are expected, particularly at the antenna's high operating  
 3733 frequency of 26 GHz where small geometric changes can have significant impacts on  
 3734 electrical properties. However, calibration through careful characterization measurements  
 3735 can be used to remove the majority of these pattern imperfections, including the relatively  
 3736 large pattern null near 45°, which will allow for the usage of the SYNCA as a test source  
 3737 for free-space CRES experiments utilizing antenna arrays. In the next section we use the  
 3738 VNA measurements obtained here as a calibration for signal reconstruction using digital  
 3739 beamforming.

### 3740 **5.3.5 Beamforming Measurements with the SYNCA**

3741 Digital beamforming is a standard technique for signal reconstruction using a phased  
 3742 array [71]. The SYNCA, since it exhibits the same cyclotron phases as a trapped electron,  
 3743 can be used to perform simulated CRES digital beamforming reconstruction experiments  
 3744 on the bench-top without the need for the magnet, cryogenics, and vacuum systems  
 3745 required by a full CRES experiment. The fields received by the individual elements  
 3746 of the antenna array will have phases dependent on the spatial position of the source  
 3747 relative to the antennas. Therefore, a simple summation of the received signals will fail  
 3748 to reconstruct the signal due to destructive interference between the individual channels  
 3749 in the array. However, applying a phase shift associated with the source's spatial position

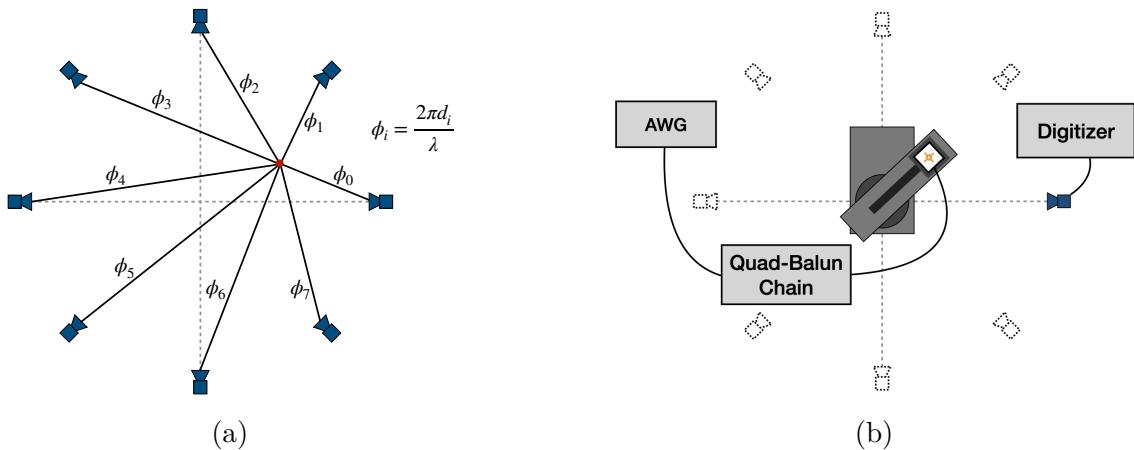


Figure 5.21: (a) A depiction of the relative phase differences for signals received by a circular antenna array from an isotropic source. The phases correspond to a unique spatial position. (b) A schematic of the setup used to perform digital beamforming.

3750 removes phase differences and results in a constructive summation of the channel signals  
 3751 (see Figure 5.21). We can summarize the digital beamforming operation succinctly using  
 3752 the following equation

$$y[t_n] = \sum_{m=0}^{N-1} x_m[t_n] A_m e^{i\phi_m}, \quad (5.35)$$

3753 where  $y[t_n]$  represents the summed array signal at time  $t_n$ ,  $x_m[t_n]$  is the signal received  
 3754 by channel  $m$  at time  $t_n$ ,  $\phi_m$  is the phase shift applied to the signal received at channel  
 3755  $m$ , and  $A_m$  is an amplitude weighting factor that accounts for the different signal power  
 3756 received by individual channels. By changing the digital beamforming phases, the point  
 3757 of constructive interference can be scanned across the sensitive region of the array to  
 3758 search for the location of a radiating source, which is identified as the point of maximum  
 3759 summed signal power above a specified threshold. The digital beamforming phases consist  
 3760 of two components,

$$\phi_m = 2\pi d_m / \lambda + \theta_m, \quad (5.36)$$

3761 where  $d_m$  is the distance from the  $m$ -th array element to the source, and  $\theta_m$  is the  
 3762 relative angle between the source position and the  $m$ -th antenna. The first component is  
 3763 the standard digital beamforming phase that corresponds to the spatial position of the  
 3764 source, and the second component is the cyclotron phase that corresponds to the relative  
 3765 azimuthal phase offset.

3766 With a small modification to the hardware used to characterize the SYNCA (see  
 3767 Figure 5.19), we can perform a digital beamforming reconstruction of a synthetic CRES  
 3768 event. By replacing the VNA with an arbitrary waveform generator (AWG), the SYNCA  
 3769 can be used to generate cyclotron radiation with an arbitrary signal structure, which  
 3770 can then be detected by digitizing the signals received by the horn antenna. Rotational  
 3771 symmetry allows us to use the rotational stage of the positioning system to rotate the  
 3772 SYNCA to recreate the signals that would have been received by a complete circular  
 3773 array of antennas.

3774 Using this setup, signals from a 60 channel circular array of equally spaced horn  
 3775 antennas were generated with the SYNCA positioned 10 mm off the central array axis,  
 3776 reconstructed using digital beamforming, and compared to Locust simulation (see Figure  
 3777 5.22). When the cyclotron spiral phases are not used, which is equivalent to setting  $\theta_m$   
 3778 in Equation 5.36 to zero, the SYNCA's position is reconstructed as a relatively faint ring  
 3779 as predicted by simulation. However, when the appropriate cyclotron phases are used  
 3780 during the beamforming procedure, both the simulated electron and the SYNCA appear

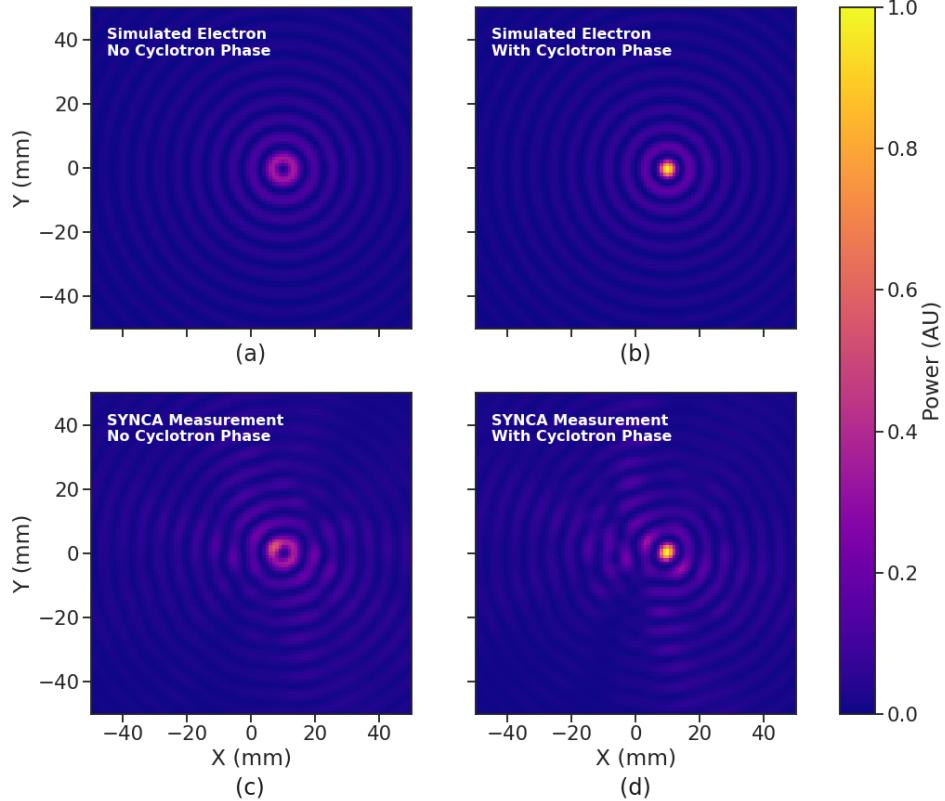


Figure 5.22: Digital beamforming maps generated using a simulated 60 channel array and electron simulated using the Locust package. (a) and (b) show the beamforming maps for simulated electrons without the cyclotron spiral phases and with the cyclotron spiral phases respectively. (c) and (d) show the beamforming maps produced from SYNCA measurements. We observe good agreement between simulated electrons and the SYNCA measurements.

as a single peak of high relative power corresponding to the source position. Therefore, we observe good agreement between the simulated and SYNCA reconstructions. While it may seem that for the case with no cyclotron phase corrections the ring reconstructs the position of the electron as effectively as beamforming with the cyclotron phase corrections, it is important to note that the simulations and measurements were generated without a realistic level of thermal noise. The larger maxima region and lower signal power, which occurs without the cyclotron phase corrections, significantly reduce the probability of detecting an electron in a realistic noise background.

To bound the beamforming capabilities of the synthetic array of horn antennas, we performed a series of beamforming reconstructions where the SYNCA was progressively moved off the central axis of the array (see Figure 5.23). To extract an estimate of the

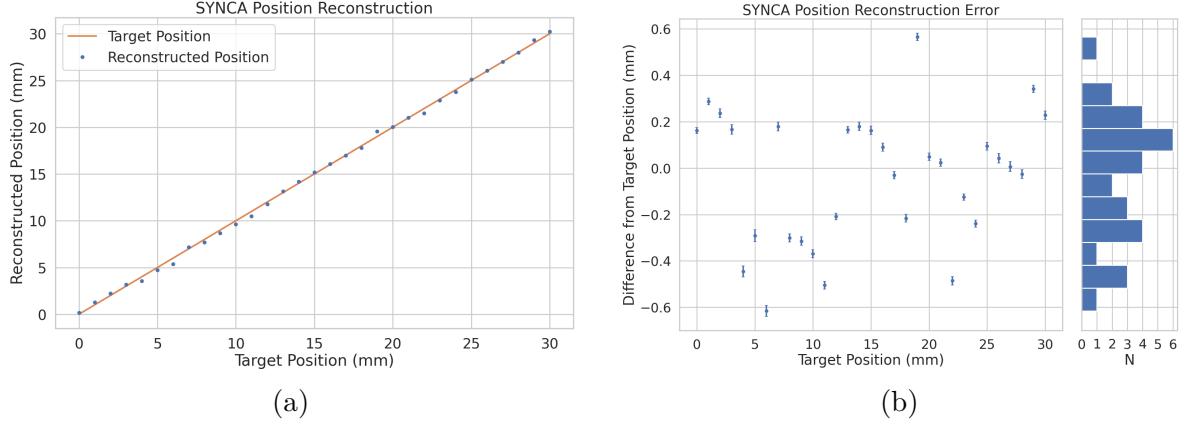


Figure 5.23: A plot of the SYNCA’s reconstructed position using the synthesized horn-antenna array and digital beamforming. (a) Shows the reconstructed position of the SYNCA compared with the target position indicated by the positioning system readout. (b) Shows the reconstruction error, which is the difference between the target and reconstructed positions. The error bars in (b) are the uncertainty in the mean position of the 2D Gaussian used to fit the digital beamforming reconstruction peak obtained from the fit covariance matrix. The mean fit position uncertainty of 0.02 mm is an order of magnitude smaller than the typical reconstruction error of 0.3 mm obtained by calculating the standard deviation of the difference between the reconstructed and target position.

position of the SYNCA using the digital beamforming image we apply a 2-dimensional (2D) Gaussian fit to the image data and extract the estimated centroid value. We find that the synthetic horn antenna array reconstructs the position of the SYNCA with a  $1\sigma$ -error of 0.3 mm with no apparent trend across the 30 mm measurement range. This reconstruction error is an order of magnitude larger than mean fit position uncertainty of 0.02 mm indicating that systematic effects related to the SYNCA positioning system could be contributing additional uncertainty to the measurements. Note that the current mean reconstruction error of 0.3 mm is a factor of 20 smaller than the full width at half maximum of the digital beamforming peak (6 mm), which could be interpreted as a naive estimate of the position reconstruction performance of this technique. Because these measurements are intended as a proof-of-principle demonstration, we do not investigate potential sources of systematic errors further; however, we expect that a similar and more thorough investigation will be performed using the Project 8 antenna array test stand, where typical reconstruction errors can be used to estimate the energy resolution limits of antenna array designs.

3807 **5.3.6 Conclusions**

3808 In this paper we have introduced the SYNCA, which is a novel antenna design that  
3809 emits radiation that mimics the unique properties of the cyclotron radiation generated by  
3810 charged particles moving in a magnetic field. The characterization measurements of the  
3811 SYNCA validated the simulated performance of the PCB crossed-dipole antenna design.  
3812 Additionally, the SYNCA was used to estimate the position reconstruction capabilities  
3813 of a synthesized array of horn antennas and experimentally reproduced the simulated  
3814 digital beamforming reconstruction of electrons.

3815 While the SYNCA performs well, there exist discrepancies in the phase and magnitude  
3816 of the radiation pattern compared to the simulated SYNCA design that are related to  
3817 the small geometric differences in the soldered connections. Future design iterations that  
3818 replace the soldered connections with a fully surface mount design could improve the  
3819 radiation pattern at the cost of some complexity and expense. Furthermore, improving  
3820 the design of the antenna PCB and mounting system would allow the antenna to be  
3821 inserted into a cryogenic and vacuum environment where in-situ antenna measurement  
3822 calibrations could be performed.

3823 The discrepancies in the radiation pattern and phases exhibited by the as-built  
3824 SYNCA should not greatly impact its performance as a calibration probe. Both magni-  
3825 tude and phase variations can be accounted by applying the SYNCA characterization  
3826 measurements as a calibration to the data collected by the antenna array test stand. The  
3827 separate calibration of the SYNCA radiation does not impact the primary goals for the  
3828 antenna array test stand which are array calibration and signal reconstruction algorithm  
3829 performance characterization, because it can be performed with standard reference horn  
3830 antennas with well understood characteristics.

3831 The SYNCA antenna technology advances the CRES technique by providing a  
3832 mechanism to characterize free-space antenna arrays for CRES measurements without  
3833 the need for a magnet and cryogenics system, which would be required for calibration  
3834 using electron sources. Both the Project 8 collaboration as well as future collaborations  
3835 which are developing antenna array based CRES experiments can make use of SYNCA  
3836 antennas as an important component of their calibration and commissioning phases.

## 3837 5.4 FSCD Antenna Array Measurements with the SYNCA

### 3838 5.4.1 Introduction

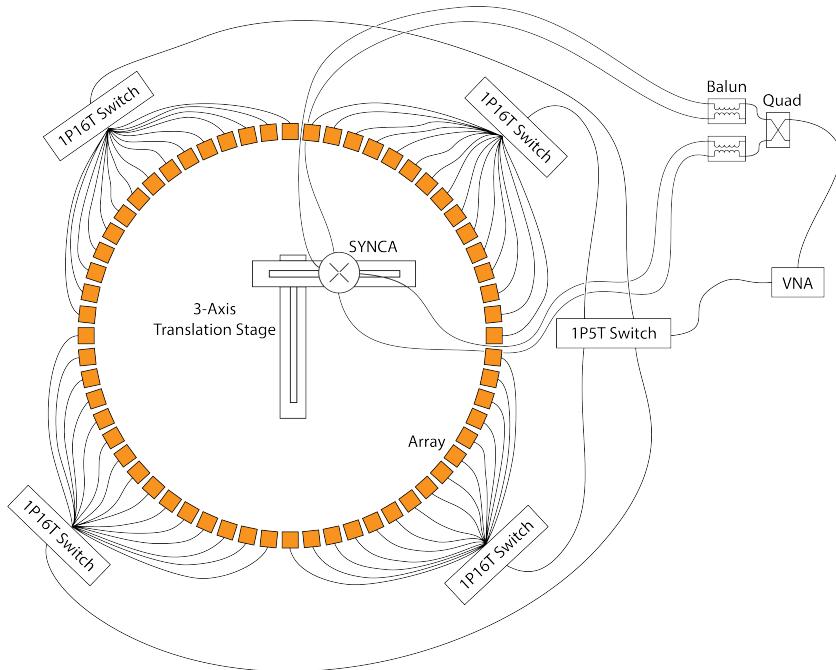


Figure 5.24: A diagram of the array measurement system used to test the prototype FSCD antenna array. A VNA is used as the primary measurement tool, which is connected to the array through a series of switches. The other port of the VNA connects to the SYNCA through the quad-balun chain used to provide the SYNCA feed signals. During measurements the SYNCA is positioned inside the center of the antenna array and translated to different radial and axial positions using a 3-axis manual translation stage setup.

3839 Using the SYNCA we can perform full-array measurements of prototype versions  
3840 of the FSCD antenna array to test its performance with a realistic cyclotron radiation  
3841 source (see Figure 5.24). The goal is to check how the measured power received by  
3842 the array compares to FSCD simulations as a function of the radial and axial position  
3843 of the SYNCA. These measurements are intended to validate the antenna research  
3844 and development by Project 8, which has been driven primarily by simulations with  
3845 Locust [25] and CREsana (see Section 4.2.3), and identify any discrepancies with these  
3846 simulations tools. This knowledge will provide confidence in the simulations necessary  
3847 for the analysis of the sensitivity of larger antenna array based CRES experiment designs  
3848 to the neutrino mass.

3849 As shown in Section 5.3, the SYNCA does have some radiation pattern imperfections  
3850 that complicate the comparison between measurement and simulation data. One way to  
3851 disentangle some of the effects of these imperfections is to perform an additional set of  
3852 measurements using a synthetic antenna array setup along with the SYNCA antenna.  
3853 Since the synthetic array setup uses only a single array antenna, the data should be  
3854 free of errors associated with individual antenna differences and multi-path interference,  
3855 which are two error sources being tested with the full-array setup. By comparing the  
3856 synthetic array data to the FSCD array data and to simulation data one can evaluate the  
3857 significance of these effects relative to the errors introduced by SYNCA imperfections.

### 3858 **5.4.2 Measurement Setups**

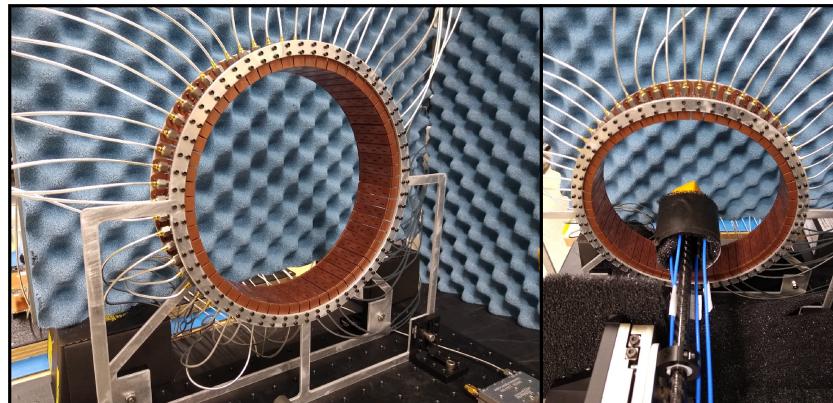
#### 3859 **5.4.2.1 FSCD Array Setup**

3860 The antenna design that composes the array is the 5-slot waveguide antenna developed  
3861 for the FSCD experiment (see Figure 5.25a). The antenna is 5 cm long and is constructed  
3862 out of WR-34 waveguide with a 2.92 mm coax connector located at the center of the  
3863 antenna. Copper flanges located on both ends of the antenna are used to mount the  
3864 antenna in the array support structure. The antennas are supported by two circular steel  
3865 brackets that can be bolted to both ends of the waveguide to construct the circular array  
3866 (see Figure 5.25b). The antenna array consists of sixty identical waveguide antennas  
3867 with a radius of 10 cm. The array is mounted perpendicular to an optical breadboard  
3868 surface using a pair of the steel brackets, which provide sufficient space for the coaxial  
3869 cable connections and allows for easy positioning of the SYNCA antenna. The SYNCA is  
3870 mounted on the end of a carbon fiber rod attached to a set of manual translation stages,  
3871 which are used to move the SYNCA antenna to different positions inside the array (see  
3872 Figure 5.25c). The stages allow for independent motion in three different axes and can  
3873 position the SYNCA at radial distances up to 5 cm from the center.

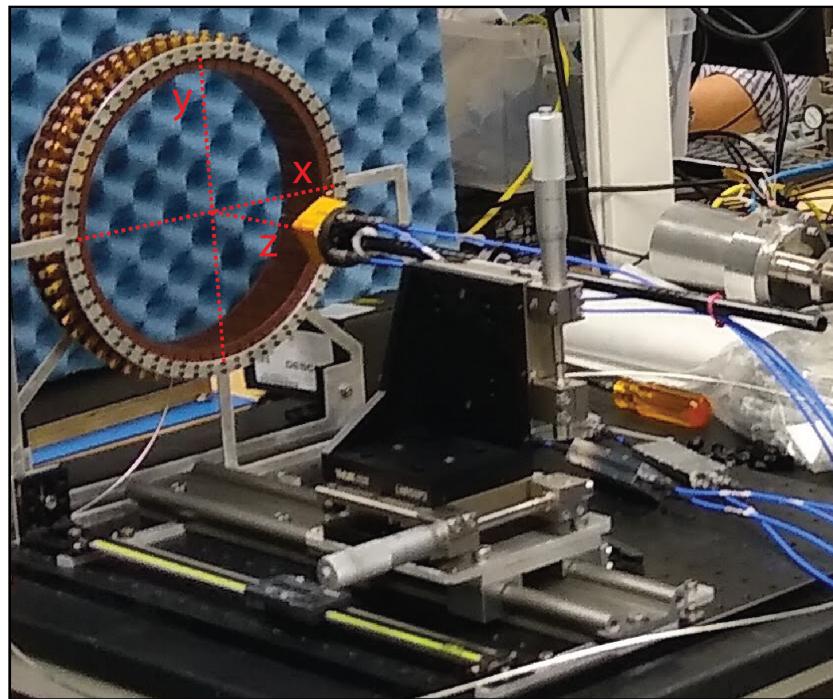
3874 Data acquisition is accomplished using a two-port VNA in combination with a series  
3875 of microwave switches that allow the VNA to connect to each channel in the array . The  
3876 first port of the VNA is connected to the quad-balun chain used to feed the SYNCA (see  
3877 Section 5.3), and the second port of the VNA connects to a 1P5T microwave switch. The  
3878 1P5T switch is connected to four separate 1P16T switch boards that connect directly  
3879 to the array. The data acquisition is controlled by a python script running on a lab  
3880 computer, which is connected to the VNA and an Arduino board programmed to control  
3881 the microwave switches. The script uses the switches to iteratively connect each of the



(a)



(b)



(c)

Figure 5.25: Photos of the prototype FSCD antenna (a), the FSCD array and SYNCA (b), and the translation stages and coordinate system used to position the SYNCA (c).

3882 antennas in the array to the VNA. The VNA is configured to load a specific calibration  
3883 file for each antenna channel and performs the measurements of all available S-parameters.  
3884 The separate calibration files is an attempt to remove phase and magnitude errors caused  
3885 by different propagation through the RF switches. Array measurements were performed  
3886 for the set of SYNCA positions consisting of radial (x-axis) positions from 0 to 50 mm in  
3887 5 mm steps and axial (z-axis) positions from 0 to 50 mm in 5 mm steps resulting in 121  
3888 array measurements. At each SYNCA position we measured the two-port S-parameter  
3889 matrix using a linear frequency sweep from 25.1 to 26.1 GHz with 101 discrete frequencies.

3890 **5.4.2.2 Synthetic Array Setup**

3891 A photograph of the setup used to perform the synthetic array measurements is shown  
3892 in Figure 5.26. One important difference between this setup and the FSCD array setup  
3893 is that the synthetic array measurements were performed with a waveform generator and  
3894 digitizer instead of a VNA. The electronics configuration is identical to the diagram in  
3895 Figure 5.7b. Despite the differences, one is still able to compare the measured phases of  
3896 the synthetic array and the relative magnitude of the power, since the digitized signal  
3897 power is directly proportional to S21.

3898 The arbitrary waveform generator in the setup is configured to produce a 64 MHz  
3899 sine wave signal that is up-converted to 25.864 GHz using a mixer and the VNA source.  
3900 This signal is passed through a bandpass filter and fed to the SYNCA quad-balun chain.  
3901 A single FSCD antenna is positioned 10 cm from the SYNCA and aligned vertically so  
3902 that the center of the 5-slot waveguide is in the plane of the SYNCA PCB (see Figure  
3903 5.26). This position corresponds to  $z = 0$  in Figure 5.25c. The SYNCA is rotated  
3904 in three degree steps to synthesize an antenna array with 120 channels. This channel  
3905 count is more than could physically fit in a 10 cm radius array, but there is no cost to  
3906 over-sampling. Additionally, over-sampling allows for a check of the smoothness of the  
3907 antenna array radiation pattern. The signals from the FSCD antenna are down-converted  
3908 using the second mixer connected to the VNA source before being digitized at 250 MHz  
3909 and saved to disk. Several synthetic array measurement scans were performed by using  
3910 the linear translation stage to change the radial position of the SYNCA. In total eight  
3911 scans were taken from 0 to 35 mm using a radial position step size of 5 mm.

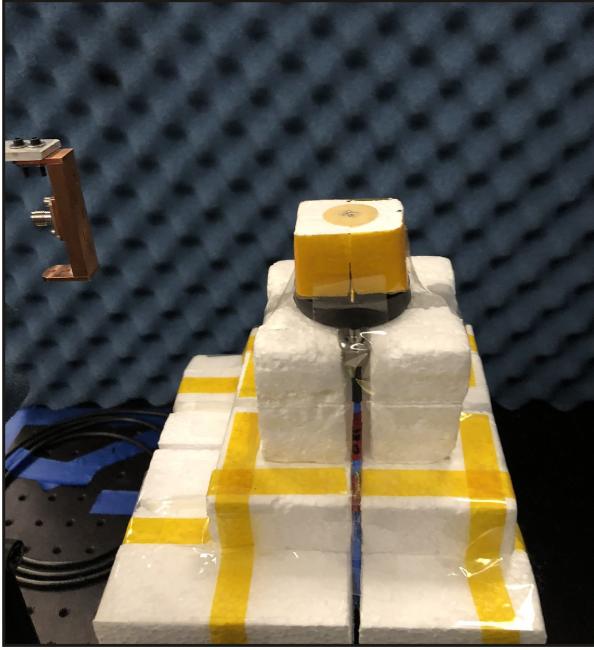


Figure 5.26: A photo of the FSCD antenna and the SYNCA in the synthetic array measurement setup at Penn State.

### **5.4.3 Simulations, Analysis, and Results**

The Locust and CRESana simulation packages utilize the antenna transfer functions to calculate the power that would be received by each antenna from a CRES electron. The equivalent quantity in the measurement setup is the S21 matrix element, which indicates the ratio of the power received by an antenna in the array to the amount of power delivered to the SYNCA. Therefore, the analysis focuses on comparing the relative magnitudes and phase of the S21 parameters measured by the VNA as a function of the array channel and the SYNCA position. Additionally, we apply a beamforming reconstruction to the S21 data to evaluate how the summed power and beamforming images change as a function of the position of the SYNCA.

#### **5.4.3.1 Simulations**

Simulations for the FSCD array measurements were performed using CRESana, which performs analytical calculations of the EM-fields produced by an electron at the position of the antennas. At each sampled time CRESana computes the electric field vector at the antenna positions, which is projected onto the antenna polarization axis to obtain the co-polar electric field. The magnitude of the co-polar electric field is then multiplied by

3928 a flat antenna transfer function to calculate the corresponding voltage signal. CRESana  
 3929 simulations exploit the flat transfer functions of the FSCD antennas, which allows the  
 3930 electric field to be multiplied by the antenna transfer function rather than performing  
 3931 the full FIR calculation. These calculations produce a voltage time-series for each of the  
 3932 antennas in the array that can be compared to the laboratory measurements.

3933 CRESana was configured to simulate a  $90^\circ$  electron in a constant background magnetic  
 3934 field of  $\approx 0.958$  T with a kinetic energy of 18.6 keV. These parameters were chosen  
 3935 in order to mimic a CRES event near the tritium beta-decay spectrum endpoint in  
 3936 the FSCD experiment. The constant background magnetic field guarantees that the  
 3937 guiding center of the electron is stationary across the duration of the simulation which is  
 3938 consistent with the SYNCA in the laboratory measurements. Simulations were performed  
 3939 with the electron's guiding center at radial positions from 0 to 45 mm in steps of 1 mm  
 3940 and axial positions from 0 to 30 mm in steps of 1 mm. The simulations generated time  
 3941 series consisting of 8192 samples at 200 MHz for the sixty channel FSCD antenna array  
 3942 geometry.

#### 3943 5.4.3.2 Phase Analysis

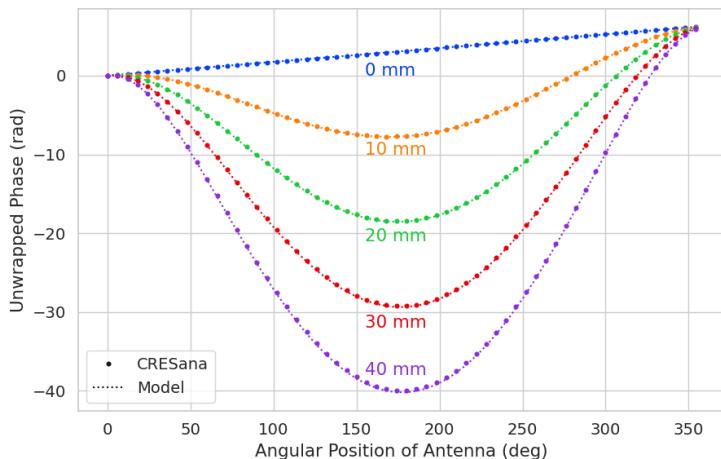


Figure 5.27: The unwrapped phases of signals received by the FSCD antenna array from an electron with a  $90^\circ$  pitch angle located in the plane of the antenna array. The data points indicated the phases extracted from simulation and the dashed lines show the model predictions.

3944 Correct modeling of the signal phases is fundamental to reconstruction for both  
 3945 beamforming and matched filter approaches. The beamforming reconstruction relies on

3946 a signal phase model developed from Locust simulations, which allows one to predict the  
3947 relative signal phases for a specific magnetic trap and electron position. The equation  
3948 for the model is

$$\phi_{ij}(t) = \frac{2\pi d_{ij}(t)}{\lambda} + \theta_{ij}(t), \quad (5.37)$$

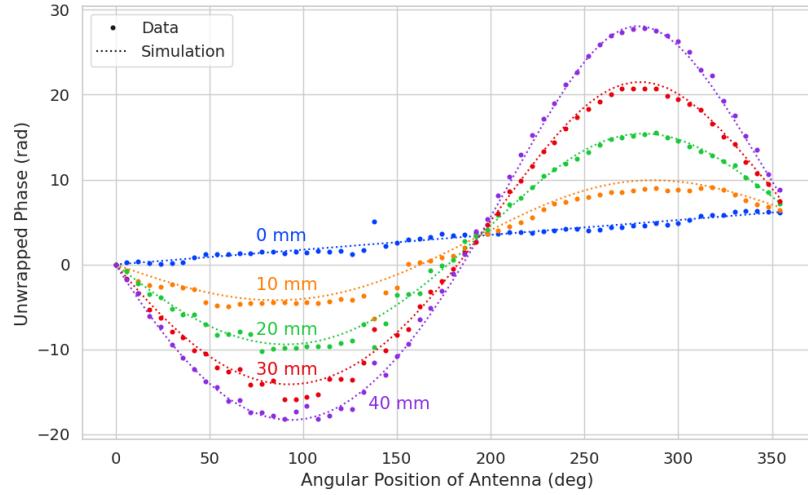
3949 where  $d_{ij}(t)$  is distance between the assumed electron position and the antenna position,  
3950 and  $\theta_{ij}(t)$  is the angular separation between the electron and antenna positions. For  
3951 details on the components of the phase model see Section 5.3.2. In Figure 5.27 we  
3952 compare the phases predicted by Equation 5.37 to phases extracted from CREsana  
3953 simulations of an electron located in the plane of the antenna array at a series of radial  
3954 positions. One observes excellent agreement between the model and simulation.

3955 The measured signal phases from the FSCD array and synthetic array are shown  
3956 in Figures 5.28a and 5.28b compared to the signal phase model. The axial position of  
3957 the SYNCA in both plots is  $z = 0$  mm, such that the plane of the PCB is aligned with  
3958 the center of the FSCD antenna. The data shown in Figure 5.28a corresponds to the  
3959 S-parameters measured at 25.80 GHz which is the frequency closest to the one used in  
3960 the synthetic array setup. The different slope and sinusoidal phases exhibited by Figure  
3961 5.28a and 5.28b reflects differences in the coordinate system for each setup. In general,  
3962 we see that the phase model predicts the large scale features of the phases quite well,  
3963 but there are some small scale deviations or errors from the phase model that do not  
3964 appear to be present in simulation.

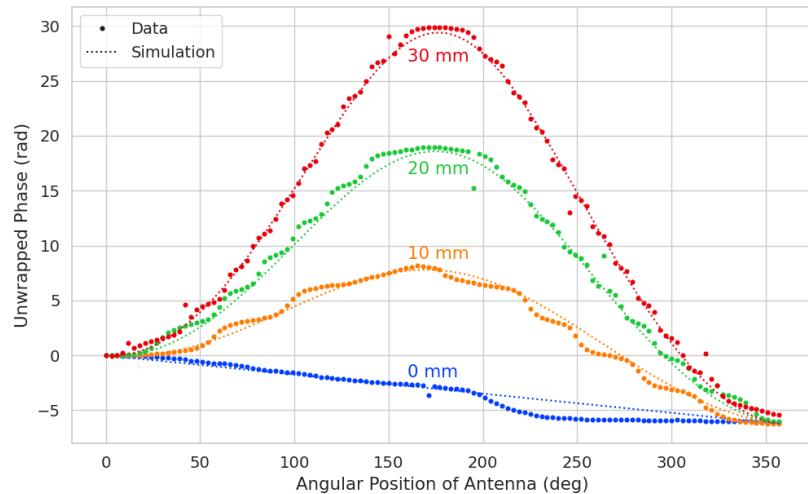
3965 A comparison of the phase errors, which are the difference between measurement and  
3966 model is shown in Figure 5.29. The FSCD array data is referred to as the JUGAAD  
3967 data in the plot legend, which is an alternative name for the FSCD array setup.

3968 The phase error at  $R = 0$  in Figure 5.29 forms a smooth curve, with the exception of  
3969 an outlier data point caused by a bug in the data acquisition script. One can attribute  
3970 the observed phase error at this position to imperfections in the antenna pattern of the  
3971 SYNCA. As the SYNCA is moved away from  $R = 0$  mm one observes that the phase  
3972 error exhibits oscillations whose frequency increases as a function of the radial position  
3973 of the SYNCA. These oscillations have the appearance of a diffraction pattern, which  
3974 is particularly clear for the radii  $\geq 15$  mm, due to the bilateral symmetry of the phase  
3975 error peaks around  $180^\circ$ .

3976 One can observe a higher average variance in the phase errors measured for the FSCD  
3977 array compared to the synthetic array. This is best seen by comparing the curves at  
3978  $R \leq 15$  mm where the smooth synthetic array curves are distinct from the relatively  
3979 noisy FSCD array errors. The extra noise in the FSCD array is most likely caused by



(a)



(b)

Figure 5.28: Plots of the measured unwrapped phases from the FSCD array (a) and the synthetic array (b) compared to the model predictions for a series of radial positions. The different phases of the sinusoidal phase oscillations in the two plots reflects differences in the coordinate systems of the measurements.

3980 differences in the radiation patterns of the antennas that make up the array as well as  
 3981 differences in the transmission lines through the switch network that introduce additional  
 3982 phase errors into the measurement. Since the synthetic array measurements use only  
 3983 a single antenna, these extra error terms are not present, which explains the relatively  
 3984 smoother phase error curves. Despite the extra phase errors in the FSCD array, it is still  
 3985 possible to observe a similar phase error oscillation effect as the SYNCA is moved away

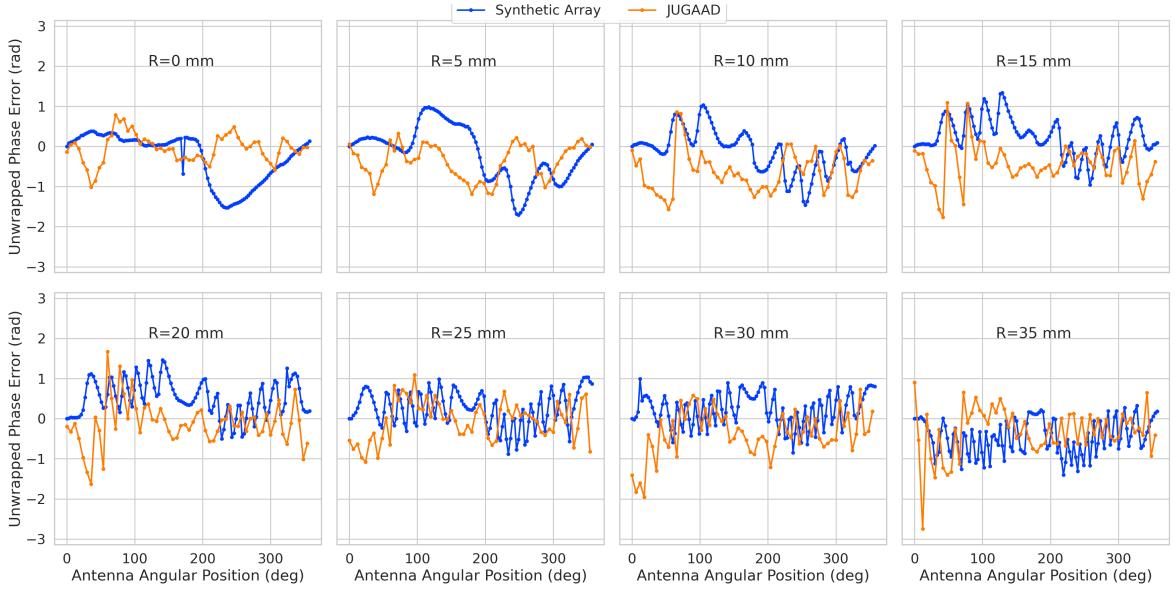


Figure 5.29: The phase errors between the measurement and model for the synthetic array (blue) and the FSCD array (orange) for a series of radial positions. The label JUGAAD refers to an alternative name for the FSCD array setup. As the SYNCA is translated off-axis phase errors with progressively higher oscillation frequency enter into the measurements.

3986 from  $R = 0$  mm.

3987 The diffraction pattern exhibited by the phase error oscillations is more easily observed  
 3988 by plotting the phase errors in a two-dimensional map, which is done in Figures 5.30a and  
 3989 5.30b. For the synthetic array ones observes a relatively clear diffraction pattern  
 3990 that emerges as the SYNCA is moved radially. The bilateral symmetry of the diffraction  
 3991 patterns is due to the bilateral symmetry of the circular synthetic array around the  
 3992 translation axis of the SYNCA. A similar pattern is also visible in the FSCD array data,  
 3993 although, it is obscured by the additional phase error that results from the multi-channel  
 3994 array.

3995 The physical origin of the phase error diffraction pattern is attributed to interference  
 3996 effects arising from path-length differences between the individual slots in the FSCD  
 3997 antenna and the SYNCA transmitter. Since we are operating in the radiative near-field of  
 3998 the FSCD antenna, the path length differences between the slots introduces a significant  
 3999 change in the summation of the signals that occurs inside the waveguide, which causes  
 4000 the radiation pattern of the antenna to change as a function of distance. Therefore, when  
 4001 the SYNCA is positioned off-axis the different path-lengths from the SYNCA to each  
 4002 antenna results in different radiation patterns leading to the observed diffraction pattern.

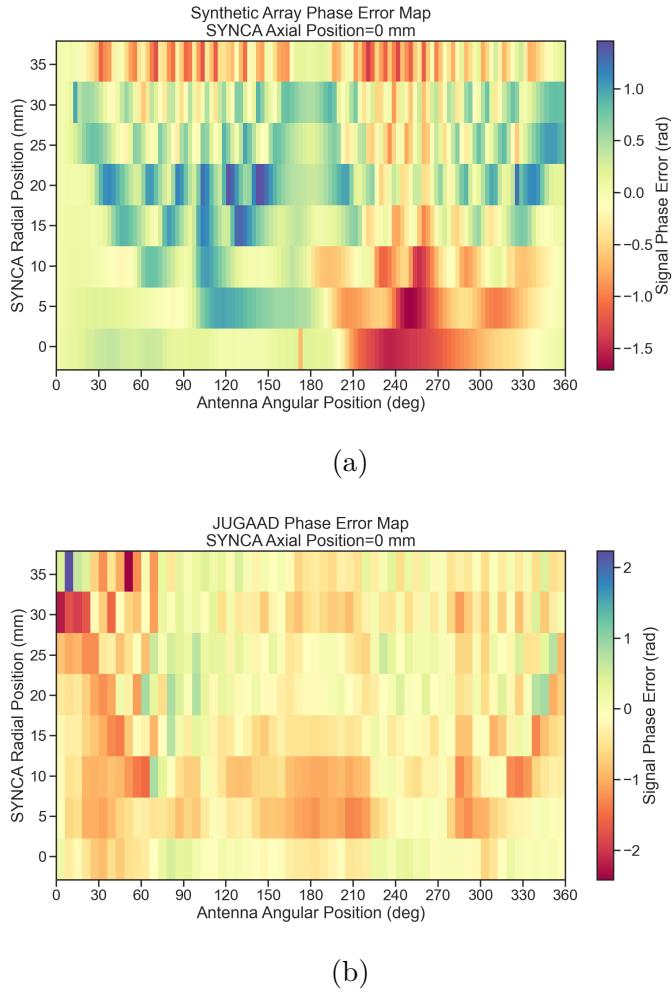


Figure 5.30: Two dimensional plots of the phase errors for the synthetic array (a) and the FSCD (JUGAAD) array (b). In both plots we observe evidence of a similar diffraction pattern with bilateral symmetry, but the FSCD array measurements have an additional phase error contribution from the different antennas and paths through the switch network.

4003 This near-field effect is not present in simulations, because in order to simplify the  
 4004 calculations we assume that the far-field approximation can be applied to the FSCD  
 4005 antennas. This means that the radiation pattern and antenna transfer functions are  
 4006 independent of the distance between the transmitter and the receiving antenna. In  
 4007 principle, we can account for these near-field effects with a more detailed simulation of  
 4008 the FSCD antennas either in CRESana or Locust, which would result in an additional  
 4009 term in the beamforming phase model. However, this would increase the computational  
 4010 intensity of the simulation software. In the next section we briefly discuss the impact of

4011 these near-field effects on the measured magnitude of the signals.

#### 4012 5.4.3.3 Magnitude Analysis

4013 Exactly as for the signal phase, one can use simulations to construct a model that  
4014 describes the magnitude of the signals received by each channel in the antenna array.  
4015 By examining the results of simulations or by analyzing the Liénard-Wiechert equation  
4016 one can show that radiation pattern from a  $90^\circ$  pitch angle electron in a magnetic field  
4017 is omni-directional. Therefore the relative magnitudes of the signals received by each  
4018 channel will be determined by the free-space power loss, which is proportional to the  
4019 inverse distance between the assumed electron position and the antenna.

4020 A consequence of this is that the signals produced in the array for electrons off the  
4021 central axis will have larger amplitudes for the antennas closer to the electron compared  
4022 to those which are further away. The amplitudes of the signals received by the array  
from an electron located at a series of radial positions are shown in Figure 5.31.

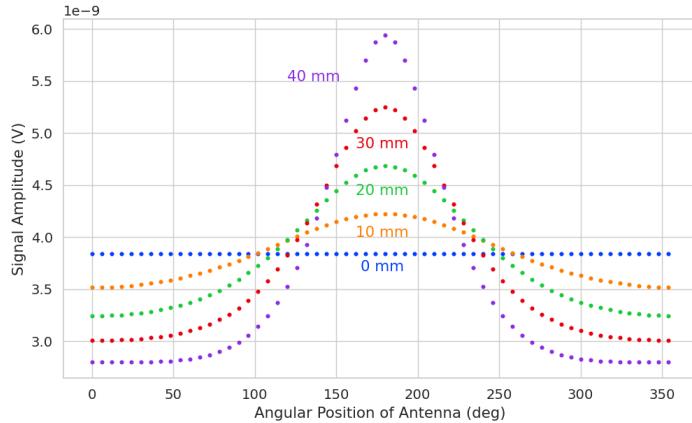


Figure 5.31: The amplitude of the signals from CREsana for the FSCD array from a  $90^\circ$  electron. As the electron is moved from  $R = 0$  the signals begin to have unequal amplitudes depending on the distance from the electron to the antenna.

4023  
4024 One expects to see a similar trend in the signal magnitudes in both the FSCD and  
4025 synthetic arrays. The normalized signal magnitudes extracted from the full and synthetic  
4026 array setups for a series of radial SYNCA positions are shown in Figure 5.32. The data  
4027 corresponds to a SYNCA axial position of  $z = 0$  mm and at a frequency 25.86 GHz. One  
4028 complication is that the radiation pattern of the SYNCA is not perfectly omni-directional,  
4029 which causes the measured magnitudes at  $R = 0$  mm to diverge from the perfectly flat  
4030 behavior exhibited by electrons.

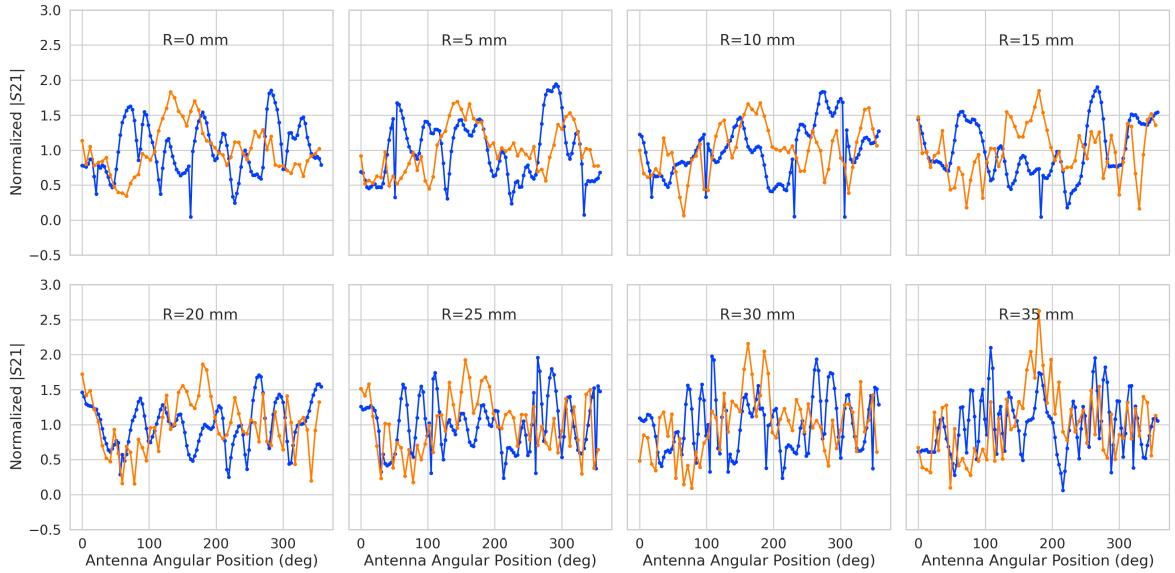
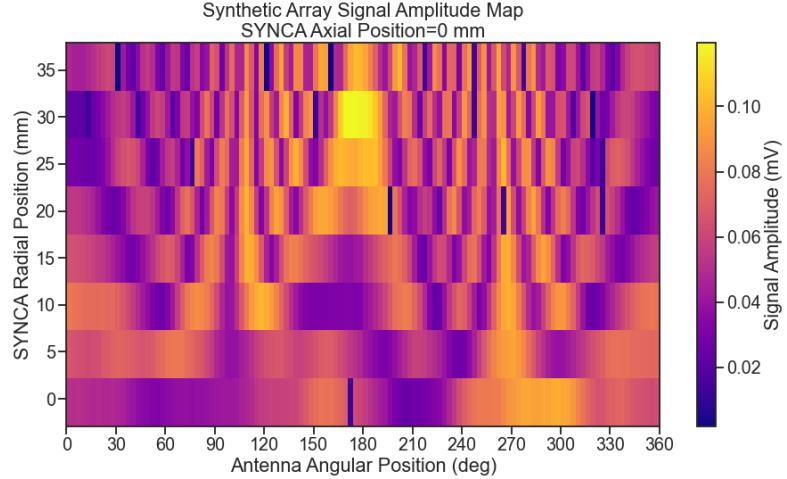


Figure 5.32: The normalized magnitudes of the S21 parameters measured in the FSCD (orange) and synthetic array (blue) setups. The dominant observed behavior as a function of radius is the increase in the number of magnitude peaks, which was noted in the phase error curves. There does not appear to be a strong change in the relative amplitude of a group of antennas as predicted by CREsana.

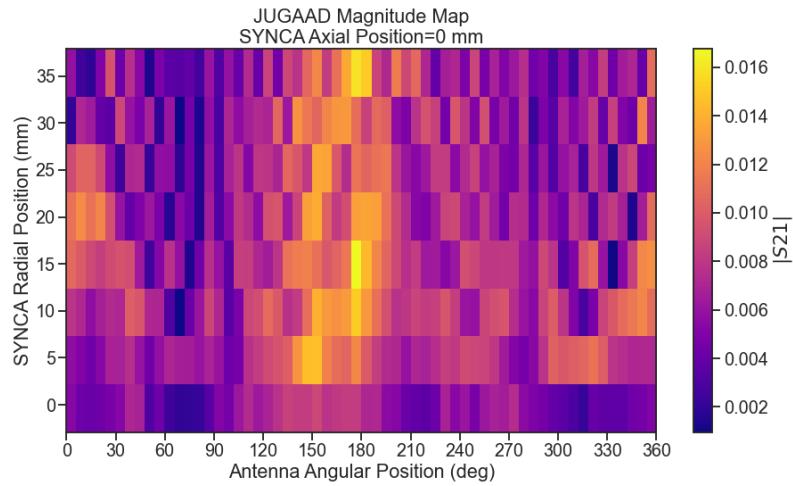
As the SYNCA is moved off-axis one observes a similar increase in the number of magnitude peaks in the synthetic array data that one would expect from a diffraction pattern, although this trend is not as stark compared to the phase data. Noticeably, there does not appear to be a set of channels with disproportionately larger amplitude at large  $R$ , which would be expected based on the trends from CREsana.

Comparing the magnitudes of the synthetic array to the FSCD array in Figure 5.32 we see that there is a similar amount of variability in the magnitudes at  $R = 0$  mm, although there is potentially more small scale error in the magnitude curve caused by channel differences in the FSCD array. We observe a similar trend in the number of magnitude error peaks in the FSCD array data to the synthetic array data, which mirrors the diffraction effect observed in the phase data. The diffraction effect can be visualized more clearly by plotting a similar two-dimensional map of the magnitudes (see Figure 5.33).

The fact that one observes a similar diffraction pattern in the signal magnitudes as a function the SYNCA position reinforces the conclusions from the phase analysis that near-field effects are having a significant impact on the radiation pattern of the FSCD array. These near-field effects lead to changes in the magnitude and phase of the



(a)



(b) The two-dimensional maps showing the diffractive pattern exhibited by the FSCD and synthetic array signal magnitudes.

Figure 5.33

4048 radiation pattern of the FSCD antenna as a function of distance. If left uncorrected these  
 4049 errors reduce detection efficiency by causing power loss in the beamforming or matched  
 4050 filter reconstruction due to phase mismatch. We explore the impact of these phase and  
 4051 magnitude errors on beamforming in the next section.

#### 4052 5.4.3.4 Beamforming Characterization

4053 Errors in the signal magnitudes and phases lead to errors in signal reconstruction. For  
 4054 example, a matched filter reconstruction requires accurate knowledge of the signals in

4055 each channel to achieve optimal performance. Uncorrected errors leads to mismatches  
 4056 between the template and signal, which reduces detection efficiency and introduces  
 4057 uncertainty in the parameter estimation. In this section, we analyze the beamformed  
 4058 signal amplitude as a function of the position of the SYNCA to quantify the impact of  
 4059 the phase and magnitude errors on signal reconstruction. Because of the imperfections  
 4060 in the SYNCA source, it is inappropriate to directly compare the beamformed signal  
 4061 amplitude of the FSCD array or synthetic array. Such a comparison would not allow  
 4062 one to disentangle losses that occur because of the antenna array from those that occur  
 4063 because of the source. Therefore, we focus on comparing the beamforming of the FSCD  
 4064 array to the synthetic array.

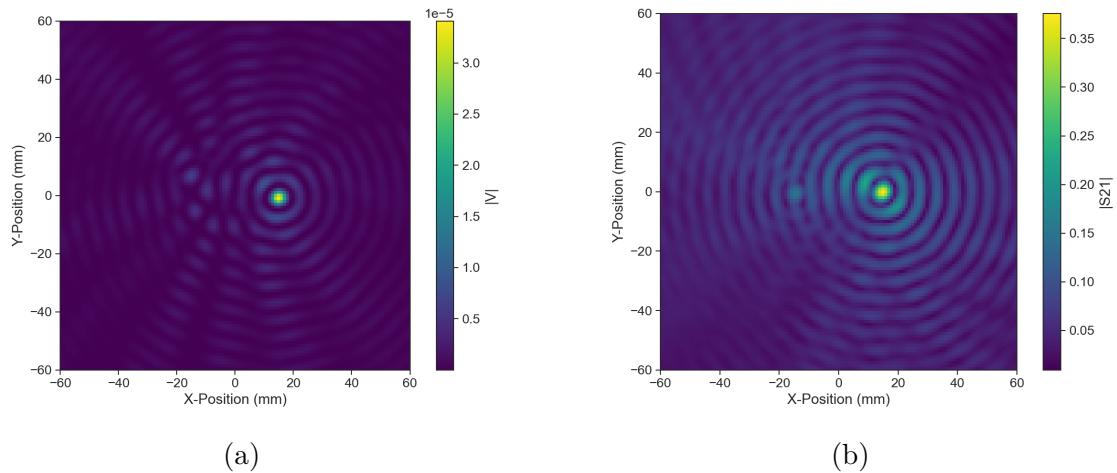


Figure 5.34: Beamforming images from the synthetic array (a) and FSCD array (b) setups with the SYNCA positioned 15 mm off the central axis. In both images we see a clear maxima that corresponds to the true SYNCA position. However, in the FSCD array there is an additional faint peak located at the opposite position of the beamforming maximum. This additional peak is the mirror of the true peak and is the result of reflections between antennas in the FSCD array.

4065 The first method of comparison is to analyze the images generated by applying the  
 4066 beamforming reconstruction specified in Section 4.3.1 to the FSCD and synthetic array  
 4067 data (see Figure 5.34). The beamforming grid consisting of a square  $121 \times 121$  grid  
 4068 spanning a range of -60-mm to 60 mm in the x and y dimensions. The beamforming  
 4069 images formed from the synthetic array produces a three-dimensional matrix where each  
 4070 grid position contains a summed time series. A single beamforming image is formed from  
 4071 this data matrix by taking the mean over the time dimension. In the case of the FSCD  
 4072 array, the VNA generates frequency domain data such that each grid position contains a  
 4073 summed frequency series produced by the VNA sweep. For this data a single image is

4074 formed by averaging in the frequency domain.

4075 There is a clear difference between the synthetic and FSCD array beamforming images,  
 4076 which is the additional faint beamforming maxima located directly opposite the maxima  
 4077 corresponding to the SYNCA position. The images in Figure 5.34 were generated with  
 4078 data collected at a SYNCA radial position of 15 mm, which agrees well with the observed  
 4079 beamforming maximum in both images. We observe that the faint beamforming peak is  
 4080 located directly opposite of the true beamforming maximum similar to a mirror image.  
 4081 Therefore, the origin of this additional feature appears to be reflections between the two  
 4082 sides of the circular antenna array that are not present for the synthetic array since only  
 4083 a single physical antenna is used.

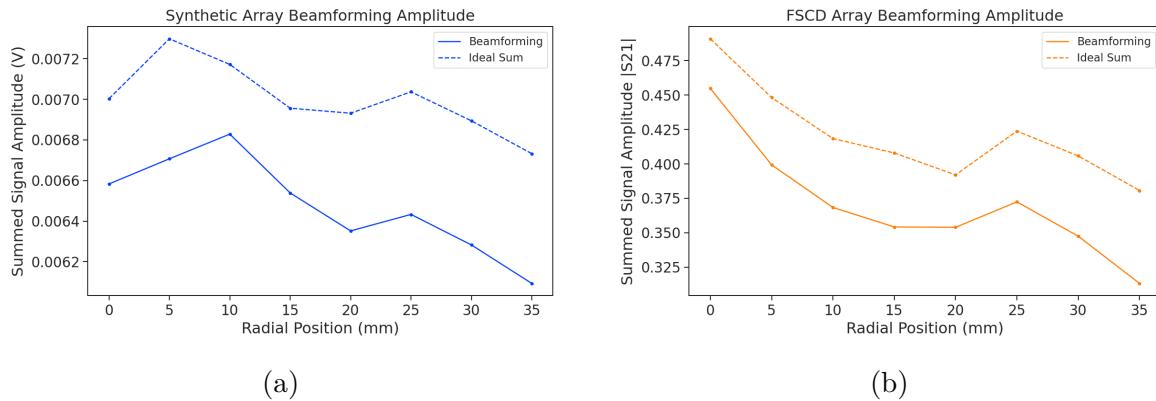


Figure 5.35: A comparison of the maximum signal amplitude obtained by beamforming to the signal amplitude obtained with an ideal summation as a function of the radial position of the SYNCA. The amplitudes for the synthetic array are shown in (a) and the FSCD array are shown in (b). In both setups we observe that the signal amplitudes obtained from beamforming are smaller than the signal amplitude that could be attained with the ideal summation without phase mismatch.

4084 From the beamforming images we extract the maximum amplitude, which we plot  
 4085 as a function of the radial position of the SYNCA (see Figure 5.35). The phase errors  
 4086 we observed in the FSCD and synthetic arrays leads to power loss at the beamforming  
 4087 stage due to phase mismatches between the signals at different channels. This power  
 4088 loss can be quantified by comparing the signal amplitude obtained from beamforming to  
 4089 the amplitude which would be obtained from an ideal summation. We perform the ideal  
 4090 summation by phase shifting each array channel to the same phase and then summing.  
 4091 The comparison between the beamforming and ideal sums is shown in Figure 5.35, where  
 4092 we observe that both the synthetic and FSCD arrays experience power losses from the  
 4093 beamforming summation.

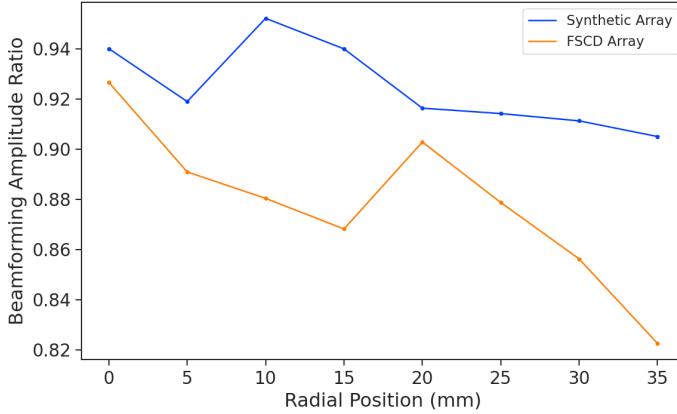


Figure 5.36: The ratio of the beamforming signal amplitude to the ideal signal amplitude for the FSCD and synthetic arrays. We see that the FSCD array has a larger power loss from phase error compare to the synthetic array which indicates that calibration errors associated with the multiple channels as well as reflections are impacting the signal reconstruction.

4094 The beamforming power loss can be quantified using the ratio of the beamforming to  
 4095 ideal signal amplitudes. Computing this ratio as a function of SYNCA radial position  
 4096 radius for the FSCD and synthetic arrays we find that the FSCD array has a uniformly  
 4097 smaller beamforming amplitude ratio, which means that the FSCD array has a larger  
 4098 beamforming power loss (see Figure 5.36). The primary contributions to the beamforming  
 4099 power loss in the synthetic array are phase errors from the SYNCA and phase errors  
 4100 from the FSCD antenna near-field. Both of these phase errors contribute to beamforming  
 4101 losses in the FSCD array, but there are clearly additional phase errors in the FSCD array  
 4102 measurements contributing to the smaller ratio. Two potential error sources include phase  
 4103 differences in the different antenna channels that could not be corrected by calibration as  
 4104 well as reflections between antennas in the array. The total effect of these additional phase  
 4105 errors is to reduce the beamforming amplitude ratio by about 5% from the beamforming  
 4106 ratio of the synthetic array. Therefore, we estimate that if no effort is made to correct  
 4107 these phase errors in an FSCD-like experiment, then we expect approximately a 10%  
 4108 total signal amplitude loss from a beamforming signal reconstruction.

#### 4109 5.4.4 Conclusions

4110 The estimated power loss of a beamforming reconstruction obtained from this analysis  
 4111 provides valuable inputs to sensitivity calculations of a FSCD-like antenna array exper-  
 4112 iment to measure the neutrino mass, since it helps to bound systematic uncertainties

4113 from the antenna array and reconstruction pipeline. This power loss lowers the estimated  
4114 detection efficiency of the experiment since some of the signal power is lost due to  
4115 improper combining between channels and also increases the uncertainty in the electron's  
4116 kinetic energy by contributing to errors in the estimation of the electron's cyclotron  
4117 frequency.

4118 If these reconstruction losses prove unacceptable there are steps that can be taken  
4119 to mitigate their effects. Some examples include the development of a more accurate  
4120 antenna simulation approach that can reproduce the observed near-field interference  
4121 patterns of the FSCD antennas and the implementation of a calibration approach that  
4122 allows for the relative phase delays of the array to be measured without changing or  
4123 disconnecting the antenna array configuration.

## **Chapter 6**

# **Development of Resonant Cavities for Large Volume CRES Measurements**

### **6.1 Introduction**

The cavity approach was originally an alternative CRES measurement technology under consideration by the Project 8 collaboration for the Phase IV experiment. After pursuing an antenna array based CRES demonstrator design for several years the increasing costs and complexity of the antenna arrays led to a reconsideration of the baseline technology for the ultimate CRES experiment planned by Project 8. Currently, a cavity based CRES experiment is the preferred technology choice for future experiments by the Project 8 collaboration including the Phase IV experiment.

In this chapter I provide a brief summary of resonant cavities and sketch out the key features of a cavity based CRES experiment. In Section 6.2 I provide a brief introduction to cylindrical resonant cavities and the solutions for the electromagnetic fields in the cavity volume.

In Section 6.3 I describe the main components of a cavity based CRES experiment, including the background and trap magnets, cavity geometry and design, and cavity coupling considerations. I also discuss some relevant trade-offs between an antenna array and cavity CRES experiment, and highlight some reasons for the transition of Project 8 to the development of a cavity based experiment.

Finally, in Sections 6.4 and 6.5, I present the design and development of an open mode-filtered cavity that could be used in a cavity based CRES experiment with atomic tritium. The results of the cavity simulations are confirmed by laboratory measurements of a proof-of-principle prototype that demonstrates key features of the design.

## 6.2 Cylindrical Resonant Cavities

Resonant cavities are sealed conductive containers, which allows us to describe the electromagnetic (EM) fields contained in the cavity volume as a superposition of resonant modes [51]. The field shapes of the resonant modes are determined by Maxwell's equations and the boundary conditions enforced by the cavity geometry. Of interest to Project 8 for CRES measurements are cylindrical cavities due to their ease of construction and integration with atom and electron trapping magnets.

### 6.2.1 General Field Solutions

Consider a long segment of conducting material with a cylindrical cross-section (see Figure 6.1). A geometry such as this can be used as a waveguide transmission line to transfer EM energy from point to point, or, if conducting shorts are inserted on both ends of the cylinder, the waveguide becomes a resonant cavity.

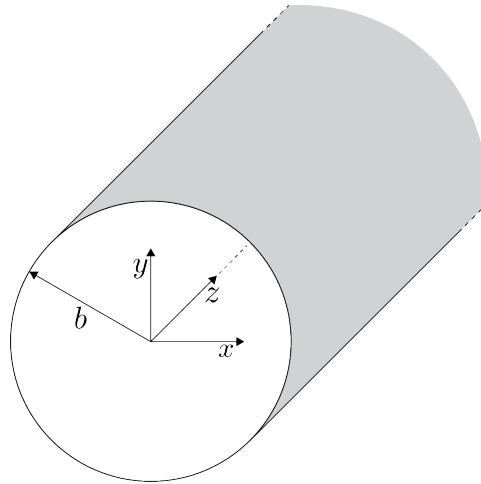


Figure 6.1: Geometry of a cylindrical waveguide with radius  $b$ .

The fields allowed inside a cylindrical cavity are determined by the boundary conditions of the cylindrical geometry. The general approach to solving the fields begins by assuming solutions to Maxwell's equations of the form

$$\mathbf{E}(x, y, z) = (\mathbf{e}(x, y) + \hat{z}e_z(x, y))e^{-i\beta z}, \quad (6.1)$$

$$\mathbf{H}(x, y, z) = (\mathbf{h}(x, y) + \hat{z}h_z(x, y))e^{-i\beta z}. \quad (6.2)$$

The solutions assume a harmonic time dependence of the form  $e^{i\omega t}$  and propagation

4164 along the positive z-axis. The functions  $\mathbf{e}(x, y)$  and  $\mathbf{h}(x, y)$  represent the transverse  
4165 ( $\hat{x}, \hat{y}$ ) components of the electric and magnetic fields respectively, and  $e_z(x, y)$ ,  $h_z(x, y)$   
4166 represent the longitudinal components. The version of Maxwell's equations in the case  
4167 where there are no source terms can be written as a pair of coupled differential equations,

$$\nabla \times \mathbf{E} = -i\omega\mu\mathbf{H}, \quad (6.3)$$

$$\nabla \times \mathbf{H} = i\omega\epsilon\mathbf{E}, \quad (6.4)$$

4168 where  $\epsilon$  and  $\mu$  are the permittivity and permeability of the material inside the waveguide  
4169 or cavity. Using the field solutions from Equations 6.1 and 6.2 one can solve for the  
4170 transverse components of the fields in terms of the longitudinal fields. Because we  
4171 are interested in cylindrical cavities it is advantageous to write the field solutions in  
4172 cylindrical coordinates. After performing this transformation the set of four equations  
4173 for the transverse field components are,

$$H_\rho = \frac{i}{k_c^2} \left( \frac{\omega\epsilon}{\rho} \frac{\partial E_z}{\partial\phi} - \beta \frac{\partial H_z}{\partial\rho} \right), \quad (6.5)$$

$$H_\phi = \frac{-i}{k_c^2} \left( \omega\epsilon \frac{\partial E_z}{\partial\rho} + \frac{\beta}{\rho} \frac{\partial H_z}{\partial\phi} \right), \quad (6.6)$$

$$E_\rho = \frac{-i}{k_c^2} \left( \beta \frac{\partial E_z}{\partial\rho} + \frac{\omega\mu}{\rho} \frac{\partial H_z}{\partial\phi} \right), \quad (6.7)$$

$$E_\phi = \frac{i}{k_c^2} \left( \frac{-\beta}{\rho} \frac{\partial E_z}{\partial\phi} + \omega\mu \frac{\partial H_z}{\partial\rho} \right), \quad (6.8)$$

4174 where  $k_c$  is the cutoff wavenumber defined by  $k_c^2 = k^2 - \beta^2$  with  $k = \omega\sqrt{\mu\epsilon}$  being the  
4175 wavenumber of the EM radiation.

4176 This set of equations can be used to solve for a variety of different modes that can be  
4177 obtained by setting conditions on  $E_z$  and  $H_z$ . For cylindrical cavities two types of modes  
4178 are allowed, which correspond to solutions where  $E_z = 0$  and  $H_z = 0$  respectively.

### 4179 6.2.2 TE and TM Modes

4180 The TE family of modes corresponds to the case where  $E_z = 0$ . This implies that  $H_z$  is  
4181 a solution to the Helmholtz wave equation

$$(\nabla^2 + k^2)H_z = 0. \quad (6.9)$$

<sup>4182</sup> For solutions of the form  $H_z(\rho, \phi, z) = h_z(\rho, \phi)e^{-i\beta z}$ , Equation 6.9 can be solved using  
<sup>4183</sup> the standard technique of separation of variables. Rather than reproduce the derivation  
<sup>4184</sup> here we shall simply quote the solutions for the transverse fields [51], which are

$$H_\rho = \frac{-i\beta}{k_{c_{nm}}} (A \sin n\phi + B \cos n\phi) J'_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.10)$$

$$H_\phi = \frac{-i\beta n}{k_{c_{nm}}^2 \rho} (A \cos n\phi - B \sin n\phi) J_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.11)$$

$$E_\rho = \frac{-i\omega\mu n}{k_{c_{nm}}^2 \rho} (A \cos n\phi - B \sin n\phi) J_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.12)$$

$$E_\phi = \frac{i\omega\mu}{k_{c_{nm}}} (A \sin n\phi + B \cos n\phi) J'_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}. \quad (6.13)$$

<sup>4185</sup> One can observe that the solutions have a periodic dependence on  $\phi$ , and radial profiles  
<sup>4186</sup> given by the Bessel functions of the first kind. The integer indices  $n$  and  $m$  arise from  
<sup>4187</sup> continuity conditions on the EM fields in the azimuthal and radial directions. For the  
<sup>4188</sup> TE modes  $n \geq 0$  and  $m \geq 1$ .  $k_{c_{nm}}$  is the cutoff wavenumber for the  $\text{TE}_{nm}$  mode given by

$$k_{c_{nm}} = \frac{p'_{nm}}{b}, \quad (6.14)$$

<sup>4189</sup> where  $b$  is the radius of the cavity or waveguide and  $p'_{nm}$  is the  $m$ -th root of the derivative  
<sup>4190</sup> of the  $n$ -th order Bessel function (see Table 6.1).

Table 6.1: A table of the values of  $p'_{nm}$ .

$n$	$p'_{n1}$	$p'_{n2}$	$p'_{n3}$
0	3.832	7.016	10.174
1	1.841	5.331	8.536
2	3.054	6.706	9.970

<sup>4191</sup> The TM mode family corresponds to the case where  $H_z = 0$ , and  $(\nabla^2 + k^2)E_z = 0$ .  
<sup>4192</sup> Again, we assume solutions of the form  $E_z(\rho, \phi, z) = e_z(\rho, \phi)e^{-i\beta z}$ , for which the general  
<sup>4193</sup> form of the solutions is the same as for the TE modes. However, the different boundary  
<sup>4194</sup> conditions for the TM modes results in particular solutions with a different from, which  
<sup>4195</sup> we shall quote here without derivation. The transverse fields of the TM modes are given  
<sup>4196</sup> by

$$H_\rho = \frac{-i\omega\epsilon n}{k_{c_{nm}}^2 \rho} (A \cos n\phi - B \sin n\phi) J_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.15)$$

$$H_\phi = \frac{-i\omega\epsilon}{k_{c_{nm}}}(A \sin n\phi + B \cos n\phi) J'_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z} \quad (6.16)$$

$$E_\rho = \frac{-i\beta}{k_{c_{nm}}}(A \sin n\phi + B \cos n\phi) J'_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.17)$$

$$E_\phi = \frac{-i\beta n}{k_{c_{nm}}^2 \rho}(A \cos n\phi - B \sin n\phi) J_n(k_{c_{nm}}\rho) e^{-i\beta_{nm}z}, \quad (6.18)$$

which one may notice are the same solutions as the TE modes with  $H$  and  $E$  flipped.

The cutoff wavenumber for the TM modes is given by,  $k_{c_{nm}} = p_{nm}/b$ , where the values of  $p_{nm}$  correspond to the  $m$ -th zero of the  $n$ -th order Bessel function (see Table 6.2).

Table 6.2: A table of the values of  $p_{nm}$ .

$n$	$p_{n1}$	$p_{n2}$	$p_{n3}$
0	2.405	5.520	8.654
1	3.832	7.016	10.174
2	5.135	8.417	11.620

### 6.2.3 Resonant Frequencies of a Cylindrical Cavity

A cylindrical cavity is constructed by taking a section of cylindrical waveguide and shorting both ends with conductive material. This means that the electric fields inside a cylindrical cavity are exactly those we derived in Section 6.2.2 with the additional condition that the electric fields must go to zero at  $z = 0$  and  $z = L$  (see Figure 6.2).

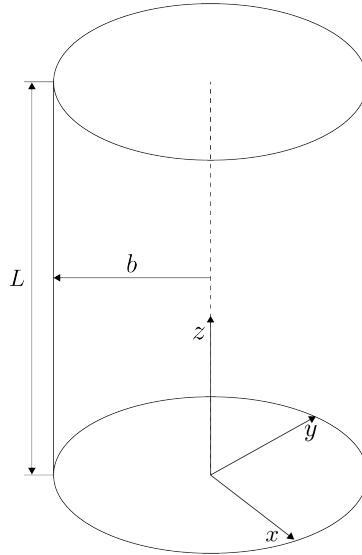


Figure 6.2: The geometry of a cylindrical cavity with length  $L$  and radius  $b$ .

4204

The transverse electric field solutions for a cylindrical waveguide are of the form

$$\mathbf{E}(\rho, \phi, z) = \mathbf{e}(\rho, \phi) (A_+ e^{-i\beta_{nm}z} + A_- e^{i\beta_{nm}z}), \quad (6.19)$$

4206 where  $A_+$  and  $A_-$  are arbitrary amplitudes of forward and backward propagating waves.

4207 In order to enforce that  $\mathbf{E}$  is zero at both ends of the cavity we require that

$$\beta_{nm}L = 2\pi\ell, \quad (6.20)$$

4208 where  $\ell = 0, 1, 2, 3, \dots$ . Using this constraint on the propagation constant we can solve

4209 for the resonant frequencies of the  $\text{TE}_{nml}$  and the  $\text{TM}_{nml}$  modes in a cylindrical cavity.

4210 For the TE modes the resonant frequencies are

$$f_{nml} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{p'_{nm}}{b}\right)^2 + \left(\frac{\ell\pi}{L}\right)^2}, \quad (6.21)$$

4211 and the frequencies of the TM modes are

$$f_{nml} = \frac{c}{2\pi\sqrt{\mu_r\epsilon_r}} \sqrt{\left(\frac{p_{nm}}{b}\right)^2 + \left(\frac{\ell\pi}{L}\right)^2}. \quad (6.22)$$

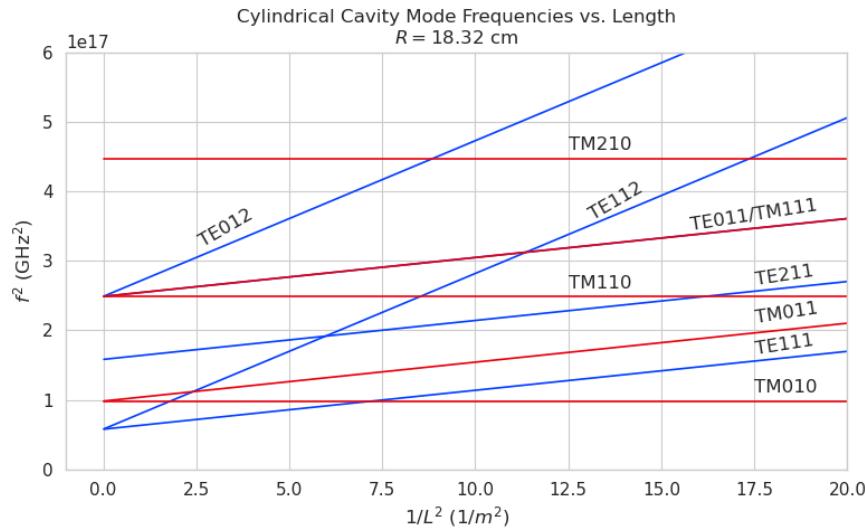


Figure 6.3: Relation of mode frequency to cavity length for a cylindrical cavity with a radius of 18.32 cm.

## 4212 6.2.4 Cavity Q-factors

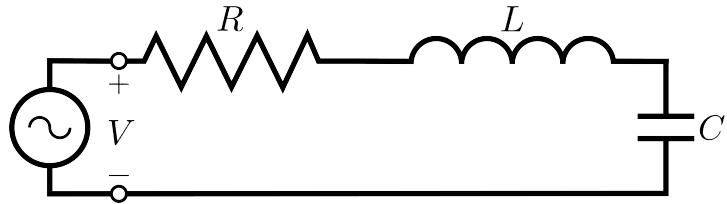


Figure 6.4: A series RLC circuit.

4213 The resonant behavior of cylindrical cavities can be modeled as a series RLC circuit  
 4214 (see figure 6.4). The input impedance of the circuit can be obtained by applying  
 4215 Kirchhoff's laws to calculate the impedance of the equivalent circuit. For a series RLC  
 4216 circuit the input impedance is

$$Z_{\text{in}} = \left( \frac{1}{R} + \frac{1}{i\omega L} + i\omega C \right). \quad (6.23)$$

4217 The resistance in the circuit represents all sources of loss in the cavity, which is primarily  
 4218 caused by the finite conductivity of the cavity walls. The inductor and capacitor represent  
 4219 the energy stored in the cavity in the form of electric and magnetic fields. If the circuit  
 4220 is being driven by an external power source we can write the input power in terms of the  
 4221 circuit input impedance and the source voltage

$$P_{\text{in}} = \frac{1}{2} Z_{\text{in}} |I|^2 = \frac{1}{2} |I|^2 \left( \frac{1}{R} + \frac{1}{i\omega L} + i\omega C \right). \quad (6.24)$$

4222 The resistor introduces a loss into the system with a power given by

$$P_{\text{loss}} = \frac{1}{2} |I|^2 R, \quad (6.25)$$

4223 and the capacitor and inductor store energies given by

$$W_e = \frac{1}{4} \frac{|I|^2}{\omega^2 C}, \quad (6.26)$$

$$W_m = \frac{1}{4} |I|^2 L, \quad (6.27)$$

4224 respectively. Using these expressions we can write the input power and input impedance

4225 expressions in terms of the lost power and stored energy

$$P_{\text{in}} = P_{\text{loss}} + 2i\omega(W_m - W_e), \quad (6.28)$$

$$Z_{\text{in}} = \frac{P_{\text{loss}} + 2i\omega(W_m - W_e)}{\frac{1}{2}|I|^2}. \quad (6.29)$$

4226 The condition for resonance in the RLC circuit is that the stored magnetic energy  
 4227 is equal to the stored electric energy ( $W_e = W_m$ ). When this occurs  $Z_{\text{in}} = R$ , which is a  
 4228 purely real impedance, and  $P_{\text{in}} = P_{\text{loss}}$ . The resonant frequency of the circuit can be  
 4229 determined from the condition  $W_e = W_m$  from which one finds that

$$\omega_0 = \frac{1}{\sqrt{LC}}. \quad (6.30)$$

4230 An important performance parameter for any resonant system is the Q-factor, which  
 4231 quantifies the quality of the resonator as the ratio of the stored energy multiplied by the  
 4232 resonant frequency to the average energy lost per second. For the series RLC circuit, the  
 4233 Q-factor is given by the expression

$$Q_0 = \omega \frac{W_e + W_m}{P_{\text{loss}}} = \frac{1}{\omega_0 RC}, \quad (6.31)$$

4234 from which one observes that as the resistance of the RLC circuit is decreased the quality  
 4235 factor of the resonator increases. From the perspective of cylindrical cavities this implies  
 4236 that as one decreases the resistance of the cavity walls it is expected that the Q-factor of  
 4237 the cavity should increase, which is indeed the case. In certain applications where a high  
 4238 Q is desireable it is possible to manufacture a cavity out of superconducting materials in  
 4239 order to minimize the power losses of the system.

4240 The Q-factor of the resonator also determines with bandwidth (BW) of the system.  
 4241 A cavity with a high Q-factor will resonant with a smaller range of frequencies than a  
 4242 cavity with a low Q-factor. To see this we can examine the behavior of the RLC circuit  
 4243 when driven by frequencies near the resonance. For a frequency  $\omega = \omega_0 + \Delta\omega$ , where  
 4244  $\Delta\omega = \omega - \omega_0 \ll \omega_0$ , we can write the input impedance as

$$Z_{\text{in}} = R + i\omega L \left( \frac{\omega^2 - \omega_0^2}{\omega^2} \right), \quad (6.32)$$

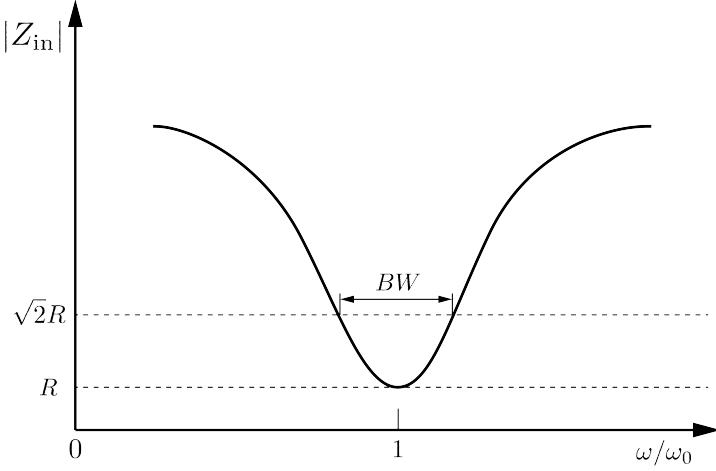


Figure 6.5: Illustration of the behavior of the input impedance of the series RLC circuit as a function of the driving frequency. The BW is proportion to the width of the resonance, which is inversely proportional to Q.

and by expanding  $(\omega^2 - \omega_0^2)/\omega^2$  to first order in  $\Delta\omega$ , we obtain

$$Z_{\text{in}} \approx R + i \frac{2RQ_0\Delta\omega}{\omega_0}. \quad (6.33)$$

Therefore, the magnitude of the input impedance near the resonance is given by

$$|Z_{\text{in}}| = R \sqrt{1 + 4Q_0^2 \frac{\Delta\omega^2}{\omega^2}}, \quad (6.34)$$

from which we observe that for the series RLC circuit the input impedance is minimized at the resonant frequency, which corresponds to the maximum input power (see Figure 6.5). The half-power BW is the range of frequencies over which the input power drops to half the input power on resonance. This occurs when  $|Z_{\text{in}}| = \sqrt{2}R$ , which corresponds to  $\Delta\omega/\omega = \text{BW}/2$ . Using Equation 6.34 one can find that

$$2R^2 = R^2(1 + Q_0^2\text{BW}^2), \quad (6.35)$$

which implies

$$\text{BW} = \frac{1}{Q_0} \quad (6.36)$$

It is important to emphasize that the Q-factor defined here,  $Q_0$ , is technically the unloaded Q. It reflects the quality of the cavity or resonant circuit without the influence of any external circuitry. In practice, however, a cavity is invariably coupled to an

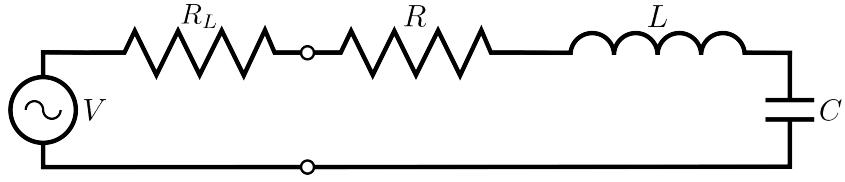


Figure 6.6: A series RLC circuit coupled to an external circuit with input impedance  $R_L$ .

4256 external circuit to drive a cavity resonance or to measure the energy of a resonant mode.  
 4257 Coupling a cavity to an external circuit changes the Q by loading the equivalent cavity  
 4258 RLC circuit (see Figure 6.6). The Q-factor of the cavity when it is loaded by an external  
 4259 circuit is called the loaded Q, which is the quantity that one actually measures when  
 4260 exciting a resonance in the cavity. Using the series RLC circuit model one can see that  
 4261 the load resistor in Figure 6.6 will add in series with the resistor in the circuit for a total  
 4262 equivalent resistance of  $R + R_L$ . Therefore, the loaded Q is given by

$$Q_L = \frac{1}{\omega_0(R + R_L)C}, \quad (6.37)$$

4263 from which one observes that the loaded Q is always less than the intrinsic Q of the  
 4264 cavity.

4265 The amount of coupling that is desireable depends on the specific application of  
 4266 the resonator. If one wants a resonator that is particular frequency selective than it  
 4267 makes sense to limit the amount of coupling to the cavity to maintain a small BW,  
 4268 alternatively, if a larger BW is need one can increase the cavity coupling by tuning the  
 4269 input impedance of the external circuit. The critical point, where maximum power is  
 4270 transferred between the cavity and the external circuit, occurs when the input impedance  
 4271 of the cavity matches the input impedance of the external transmission line. For the  
 4272 series RLC circuit on resonance, this matching condition corresponds to

$$Z_0 = Z_{in} = R, \quad (6.38)$$

4273 where  $Z_0$  is the impedance of the transmission line. The loaded Q at this critical point  
 4274 is, therefore,

$$Q_L = \frac{1}{2\omega_0 Z_0 C} = \frac{Q_0}{2}. \quad (6.39)$$

4275 One can described the degree of coupling between the cavity and an external circuit by

4276 defining a coupling factor,  $g$ , such that,

$$g = \frac{Q_0}{Q_L} - 1. \quad (6.40)$$

4277 When  $g = 1$  then  $Q_L = Q_0/2$ , and the cavity is said to be critically coupled as we  
4278 described. If  $Q_L < Q_0/2$ , then the cavity is undercoupled to the transmission line,  
4279 corresponding to  $g < 1$ . Alternatively, if  $Q_L > Q_0/2$ , then  $g > 1$ , and the cavity is  
4280 overcoupled to the transmission line. Various specialized circuits can be used to tune the  
4281 input impedance of the external circuit as seen by the cavity to achieve a wide range of  
4282 different coupling factors based on the desired application of the cavity.

## 4283 6.3 The Cavity Approach to CRES

### 4284 6.3.1 A Sketch of a Molecular Tritium Cavity CRES Experiment

4285 Resonant cavities can be used to perform CRES measurements, and they represent the  
4286 current preferred technology by the Project 8 collaboration. The basic approach to a  
4287 neutrino mass measurement using a resonant cavity and molecular tritium beta-decay  
source is illustrated by Figure 6.7.

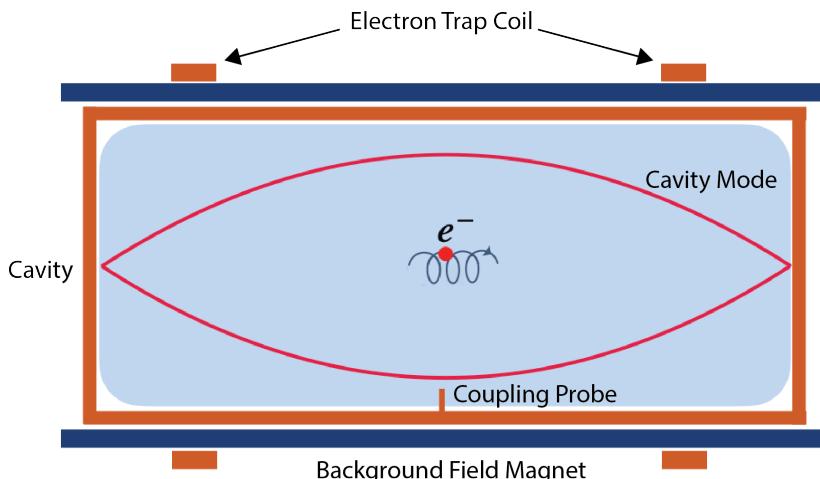


Figure 6.7: A cartoon depiction of a cavity CRES experiment. A metallic cavity filled with tritium gas is inserted into a uniform background magnetic field to perform CRES measurements. Electrons from beta-decays inside the cavity can be trapped and used to excite a resonant mode(s). By coupling to the cavity mode with a suitable probe one can measure the cyclotron frequency of the electron and perform CRES.

4288

4289 At the core of the experiment is a large resonant cavity filled with tritium gas. The  
4290 filled cavity is then placed in a uniform magnetic field provided by a primary magnet  
4291 that provides the background magnetic field. The value of the background magnetic field  
4292 sets the range of cyclotron frequencies for electrons emitted near the tritium spectrum  
4293 endpoint. When a beta-decay electron is produced in the cavity it is trapped using a set  
4294 of magnetic pinch coils that keep electrons inside the cavity volume.

4295 Electrons trapped inside the cavity do not radiate in the same way as electrons  
4296 in free-space. Effectively, the same boundary conditions that were used to derive the  
4297 resonant modes of a cylindrical cavity in Section 6.2 apply to the radiation of the electron  
4298 as well. The coupling of an electron performing cyclotron motion in a cavity has been  
4299 studied in detail for measurements of the electron’s magnetic moment [72–74] If an  
4300 electron is emitted with a kinetic energy that corresponds to a cyclotron frequency that  
4301 matches a resonant frequency of the cavity, then energy radiated by the electron excites  
4302 a corresponding resonance in the cavity. The strength of the electron’s coupling to the  
4303 cavity is given to first order by the dot product between the electrons trajectory and  
4304 the electric field vector of the resonant mode. Additional effects, such as the Purcell  
4305 enhancement [75], alter the emitted power from the free-space Larmor equation [13]. If an  
4306 electron is moving with a cyclotron frequency that is far from any resonant modes in the  
4307 cavity, then radiation from the electron is suppressed. One can interpret this somewhat  
4308 surprising effect as the metallic walls of the cavity reflecting the radiated energy back to  
4309 the electron.

4310 Detecting an electron in the cavity is accomplished by coupling the cavity to an  
4311 external transmission line that leads to an amplifier and RF receiver chain [76]. The  
4312 coupling of the cavity resonance to the amplifier occurs through a coupling probe or  
4313 aperture designed to read-out the excitation of the mode(s) excited by the electron. For  
4314 CRES measurements, the placement of a wire antenna coupling probe inside the cavity  
4315 volume leads to unacceptable losses of tritium atoms due to recombination to molecular  
4316 tritium on the antenna surface, therefore, apertures are the preferred coupling method  
4317 for cavity CRES experiments.

4318 One of the attractive features of the CRES technique for neutrino mass measurement  
4319 is the gain in statistics that comes from the differential nature of the tritium spectrum  
4320 measurement. Initially, this seems incompatible with cavities, due to the narrow reso-  
4321 nances of cavity modes giving relatively small bandwidth. However, by intentionally  
4322 over-coupling to a single cavity mode one can achieve bandwidths of a few 10’s of MHz  
4323 (see Section 6.2), which is sufficient for a measurement of the tritium spectrum endpoint

4324 region.

### 4325 **6.3.2 Magnetic Field, Cavity Geometry, and Resonant Modes**

#### 4326 **Magnetic Field and Volume Scaling**

4327 For a CRES experiment, cylindrical cavities are a natural choice since they match  
4328 the geometry of standard solenoid magnets, which are needed in order to produce the  
4329 background magnetic field for CRES measurements. Furthermore, the cylindrical shape is  
4330 compatible with a Halbach array, which is the leading choice of atom trapping technology  
4331 for future atomic tritium experiments by the Project 8 collaboration. Cylindrical  
4332 cavities also benefit from well-established machining practices that are able to achieve  
4333 high geometric precision at large lengths scales. More exotic cavity designs are under-  
4334 consideration and there are on-going efforts to investigate the potential advantages these  
4335 may have over the standard cylindrical geometry.

4336 As we saw in Section 6.2, the physical dimensions of the cavity are directly coupled  
4337 to the resonant frequencies of the cavity. This dependency links the size of the cavity to  
4338 the magnitude of the background magnetic field, because the magnetic field determines  
4339 the cyclotron frequencies of trapped electrons. Specifically, as the size of the cavity is  
4340 increased to accommodate larger volumes of tritium gas, the frequencies of the resonant  
4341 modes decrease proportionally. This requires that the magnetic field also decrease in  
4342 order to maintain coupling between electrons and the desired cavity mode.

4343 The required cavity size is ultimately determined by the required statistics in the  
4344 tritium spectrum endpoint region. Because the gas density must be kept below a certain  
4345 level to ensure that electrons have sufficient time to radiate before scattering, larger  
4346 volumes become the only way to achieve higher event statistics. To achieve the sensitivity  
4347 goals of Phase III and IV cavity volumes on the order of several cubic-meters are required,  
4348 which pushes one towards frequencies in the range of 100's of MHz.

#### 4349 **Single-mode Cavity CRES**

4350 It is tempting to consider maintaining a high magnetic field, while still increasing the size  
4351 of the cavity, in order to increase the radiated power from trapped electrons for better  
4352 SNR. However, if one were to maintain the same magnetic field while increasing the  
4353 size of the cavity, the electrons would begin to couple to higher order modes with more  
4354 complicated transverse geometries. The danger with this approach is that a complicated  
4355 mode structure could introduce systematic errors into the CRES signals. Example

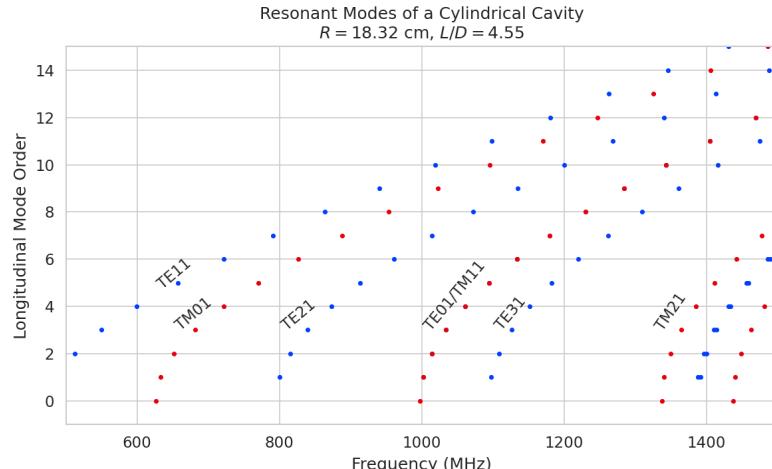
4356 systematics include unpredicted mode hybridization or changes in the mode shapes from  
4357 imperfections in the cavity construction, which would prevent reconstruction of the  
4358 electron's starting kinetic energies with adequate resolution. For this reason, it is ideal  
4359 to operate with magnetic fields that give cyclotron frequencies near the fundamental  
4360 frequency of the cavity, where the mode structure is relatively simple (see Figure 6.8).  
4361 In this frequency region it is possible to perform CRES by coupling to only a single  
4362 resonant mode, however, it is currently an open question if a single mode measurement  
4363 will provide enough information about an individual electron's position to reconstruct  
4364 the full event. Regardless, developing a solid understanding of the CRES phenomenology  
4365 when an electron is coupling to a single mode will be a necessary step towards a future  
4366 multi-mode cavity experiment.

#### 4367 Considerations for Resonant Mode Selection

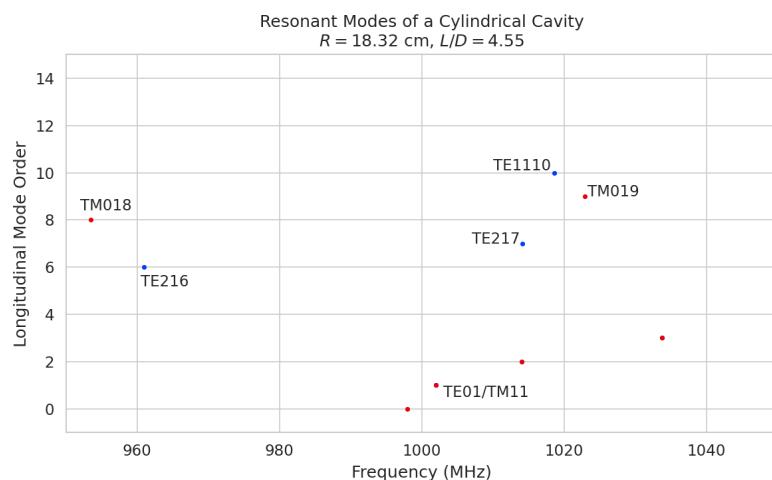
4368 A single-mode cavity experiment begs the question, which resonant mode is best for  
4369 CRES measurements? There is an immediate bias towards low order  $TE_{nm}$  and  $TM_{nm}$   
4370 modes due to the multi-mode considerations discussed above. Additionally, there is a  
4371 preference towards modes with longitudinal index  $\ell = 1$  with a single antinode along the  
4372 vertical axis of the cylindrical cavity. The reason for this is that there is a phase change  
4373 in the electric fields between antinodes that leads to modulation effects that destroy the  
4374 carrier frequency signal information.

4375 A second consideration for mode selection is the volumetric efficiency of the mode.  
4376 Volumetric efficiency can be thought of as an integral over the volume of the cavity  
4377 weighted by the relative amplitude of the mode. From the perspective of simply maximiz-  
4378 ing the volume useable for CRES measurements this integral would be as close to unity  
4379 as possible. However, there is a requirement to reconstruct the position of the electrons  
4380 inside the cavity volume so that the local magnetic fields can be used to convert the  
4381 measured cyclotron frequency to a kinetic energy. With a single mode this necessarily  
4382 requires a variable transverse mode amplitude, which lowers the volumetric efficiency, so  
4383 that position of the electron in the cavity can be estimated from the average amplitude  
4384 of the CRES signal. Longitudinal indices of  $\ell = 1$  have an advantage in volumetric  
4385 efficiency over higher order  $\ell$  modes, since there are only two longitudinal nodes, one at  
4386 each end of the cavity. Therefore, the average coupling strength of trapped electrons as  
4387 they oscillate axially is higher for  $\ell = 1$  modes.

4388 The longitudinal variation in the mode strength is ultimately critical for achieving the  
4389 energy resolution required for neutrino mass measurements. Correcting for the change in



(a)



(b)

Figure 6.8: Examples of the resonant mode frequencies of a cylindrical cavity. This cavity has a radius of 18.32 cm and a length to diameter ratio of 4.55.

4390 the average magnetic fields experienced by electrons with different pitch angles requires  
4391 that information on the axial motion of the electron be encoded into the CRES signal.  
4392 The longitudinal variation in the mode amplitude leads to amplitude modulation of the  
4393 CRES signal with a frequency proportional to the electron's pitch angle.

4394 An additional factor for mode selection is the intrinsic or unloaded Q of the mode. In  
4395 terms of SNR it is advantageous to use a mode with a very high  $Q_0$ , which is then highly  
4396 overcoupled to achieve the necessary bandwidth to cover the tritium endpoint spectrum.  
4397 This scheme leads to a decoupling of the physical cavity temperature from the effective  
4398 noise temperature after the amplifier, which allows us to achieve adequate SNR without

4399 the requirement of cooling the entire cavity to single Kelvin temperatures.

4400 An example of a resonant mode that exhibits these traits is the TE<sub>011</sub> mode. At present  
4401 the TE<sub>011</sub> mode is the preferred resonance for a single-mode cavity CRES experiment  
4402 by the Project 8 collaboration. TE<sub>011</sub> is a low order mode located in a region relatively  
4403 far from other cavity modes. Furthermore, the separation of the TE<sub>011</sub> mode can be  
4404 improved by various mode-filtering techniques discussed in Section 6.4.2 below. TE<sub>011</sub>  
4405 consists of a single longitudinal antinode that can provide pitch angle information in the  
4406 form of amplitude modulation, and has an electric field with a radial profile given by the  
4407  $J'_0$  Bessel function allowing for radial position estimation. Lastly, the TE<sub>011</sub> mode has a  
4408 relatively high intrinsic Q compared to nearby modes, which helps with SNR. Unloaded  
4409 Q's greater than 80000 are achievable for a 1 GHz TE<sub>011</sub> resonance using a copper walled  
4410 cavity.

#### 4411 **6.3.3 Trade-offs Between the Antenna and Cavity Approaches**

4412 The choice between cavities and antennas for large-scale CRES measurements is not  
4413 without trade-offs. Both the antenna array and cavity approaches are relatively immature  
4414 techniques, at present there are no known obstacles that would prevent either approach  
4415 from being used for a large scale neutrino mass experiment. The preference for cavities  
4416 is largely driven by important practical considerations that could make a cavity based  
4417 experiment significantly cheaper than an antenna experiment of similar size and scope.  
4418 However, the switch to cavities also introduces new challenges less relevant to the  
4419 antenna array, which must be solved in order for Project 8 to achieve its neutrino mass  
4420 measurement goals.

4421 One of the major relative drawbacks of the antenna array approach is the size and  
4422 complexity of the data-acquisition system. A large-scale antenna array experiment  
4423 requires  $O(100)$  antennas independently digitized at rates of  $O(10)$  to  $O(100)$  MHz. Since  
4424 there is insufficient information in a single antenna channel to detect or reconstruct the  
4425 CRES signal, the entire array output must be processed during the signal reconstruction.  
4426 Because data storage becomes an issue with these data volumes, there is a real-time  
4427 signal reconstruction requirement that allows one to detect CRES signals buried in the  
4428 thermal noise. As we discuss in Section 4.4, the computational cost of these real-time  
4429 detection algorithms are potentially quite large for even a small scale antenna array  
4430 experiment. However, the operating principle of a cavity experiment allows the CRES  
4431 signal to be detected using only a single read-out channel digitized at rates of  $O(10)$  MHz,  
4432 which reduces the cost of the data acquisition system by many orders of magnitude.

4433 From an engineering perspective, the simple geometry and thin-walls of a cylindrical  
4434 cavity are simpler to interface with the cryogenic and magnetic subsystems needed for a  
4435 CRES experiment. Whereas, the antenna array requires careful design and engineering  
4436 to accommodate the antenna array and receiver electronics in proximity to the trapping  
4437 magnets. Additionally, due to near-field interference effects, the antenna array is unable  
4438 to reconstruct CRES events within the reactive near-field distance of the antennas.  
4439 Because atom trapping requirements require magnetic fields which correspond to cyclotron  
4440 frequencies for endpoint electrons less than 1 GHz, the required stand-off distance leads to  
4441 a significant loss in useable experiment volume, necessitating larger and more expensive  
4442 magnets.

4443 Another advantage to the cavity approach is the relatively compact sideband structure,  
4444 which is a result of the low modulation index for cavity CRES signals. The axial motion  
4445 in an antenna array experiment leads to frequency modulation and sidebands. The shape  
4446 of the sideband structure is determined by the modulation index,  $h = \frac{\Delta f}{f_a}$ , where  $\Delta f$   
4447 is the size of the frequency deviation and  $f_a$  is the axial frequency. The large electron  
4448 traps required for a cubic-meter-scale experiment leads to high modulation indices, which  
4449 causes the signal spectrum to be made up of numerous low power sidebands that make  
4450 reconstruction and detection challenging. This behavior was observed in simulations  
4451 of the FSCD in which carrier power decreased with pitch angle due to the increase in  
4452 modulation index (see Figure 4.31). For cavities, however, the modulation index remains  
4453 near  $h = 1$  even for very long magnetic traps due to the high phase velocity in cavities  
4454 relative to the axial velocity of the electron. This results in an almost ideal spectrum  
4455 shape that has a strong carrier frequency with a few sidebands whose relative amplitudes  
4456 encode pitch angle information.

4457 A downside of the cavity approach is the apparent difficulty of estimating the position  
4458 of the electron using only the coupling of the electron to a single mode. The amplitude of  
4459 the TE<sub>011</sub> mode is completely independent of the azimuthal coordinate, therefore, position  
4460 reconstruction using the TE<sub>011</sub> mode is only able to estimate the radial position of the  
4461 electron. This position degeneracy may lead to magnetic field uniformity requirements  
4462 that are too challenging to meet due to mechanical uncertainties in cavity and magnet  
4463 construction, as well as uncertainties caused by nuisance external magnetic fields such  
4464 as the Earth's field and magnetic fields from building materials. A multi-mode cavity  
4465 experiment may provide a way to extract more precise information on the position of  
4466 the electron by analyzing the coupling of the electron to several modes that overlap in  
4467 different ways.

## 4468 6.4 Single-mode Resonant Cavity Design and Simulations

4469 The single-mode cylindrical cavities envisioned for the Phase III and IV experiments must  
4470 be carefully engineered in order to measure the neutrino mass with the desired sensitivity.  
4471 In this section I summarize some simulation studies performed to analyze early design  
4472 concepts for a single-mode cavity. The primary tool for these investigations was Ansys  
4473 HFSS, which was also used for the development of the SYNCA antenna described in  
4474 Section 5.3.

### 4475 6.4.1 Open Cylindrical Cavities with Coaxial Terminations

#### 4476 Design Concept

4477 A basic cavity design question relevant to Project 8’s ultimate goal of an atomic tritium  
4478 CRES experiment is how to build a cavity that can be efficiently filled with atomic  
4479 tritium. To keep the rate of atom loss from recombination on surfaces it is ideal if the  
4480 ends of the cylindrical cavity are as open as possible so that tritium atoms can flow  
4481 inside unimpeded. Additionally, one of the primary calibration techniques planned for  
4482 future CRES experiments involves CRES measurements using electrons injected from  
4483 an electron gun source, which also requires an opening at the cavity end. Cylindrical  
4484 cavities with open ends can be manufactured, however, the intrinsic Q-factors of these  
4485 cavities are orders of magnitude less than their sealed counterparts, which reduces the  
4486 signal-to-noise ratio when that cavity is used for CRES measurement.

4487 Cylindrical cavities with mostly open ends that also exhibit Q values for the  $TE_{01\ell}$   
4488 modes similar to sealed cavities can be built by using coaxial endcaps to terminate the  
4489 cavity. Cavities of this type have been manufactured for specialized applications related  
4490 to the measurements of the dielectric constants of liquefied gasses (see Figure 6.9) [77, 78].  
4491 This cavity design leaves the ends of the cavity wide open, but retains high Q-values for  
4492 the  $TE_{01\ell}$  modes due to the coaxial endcap, which are designed to perfectly reflect the  
4493 electric fields of  $TE_{01\ell}$  modes. Coupling to the  $TE_{01\ell}$  mode is achieved via an aperture  
4494 located at the center of the cavity wall.

4495 A cavity similar to Figure 6.9 is a candidate design for the future CRES experiments  
4496 by Project 8, since it appears to elegantly solve many practical issues that arise when  
4497 combining cavity CRES and atomic tritium. The coaxial endcaps leave significant regions  
4498 of the cavity ends completely open, which allows for the entrance of atomic tritium as  
4499 well as the pumping away of molecular tritium that has recombined on the cavity walls.

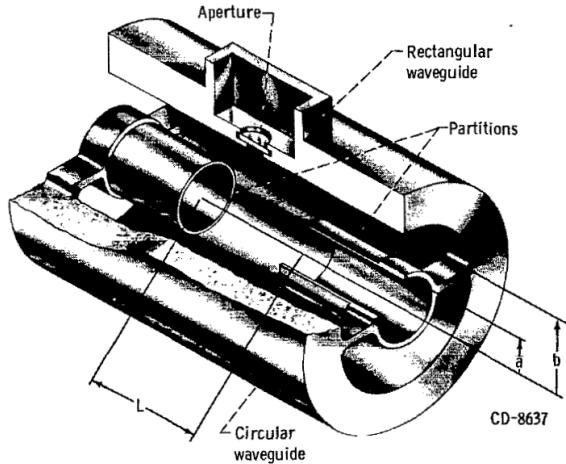


Figure 6.9: An image of an open cavity with coaxial terminations used for dielectric constant measurements. Figure from [78].

4500 These open ends are achieved while preserving the high Q-values of the  $TE_{01\ell}$  modes,  
 4501 which is important for extracting as much signal power from the electron as possible. In  
 4502 subsequent sections we shall analyze this cavity design in more detail, primarily by using  
 4503 HFSS simulations to analyze the resonant mode structure of this cavity geometry.

#### 4504 **Coaxial Terminator Constraints**

4505 The reason that coaxial endcaps can be used to achieve high Q-values for the  $TE_{01\ell}$   
 4506 modes is that the electric fields for these modes are purely azimuthally polarized (see  
 4507 Equations 6.12 and 6.13). Therefore, the boundary conditions that require the electric  
 4508 field to go to zero at the cavity ends can be supplied using a coaxial partition of the  
 4509 correct radius (see Figure 6.10). Because the cylindrical shape enforced by the partition  
 4510 does not match the boundary conditions of other cavity modes, these terminations also  
 4511 significantly suppress the Q-factors of non- $TE_{01\ell}$  modes, which is potentially beneficial  
 4512 for a single-mode cavity CRES experiment.

4513 The correct radius of the cylindrical partition is derived by setting up the boundary  
 4514 value problem in Figure 6.10, and analyzing the reflection and transmission coefficients  
 4515 for waves incident on the coaxial terminators. The basic problem is to identify the radius  
 4516  $a$  where the reflection coefficient for the  $TE_{01\ell}$  modes becomes equal to 1. One can show  
 4517 that if the coaxial partitions are made sufficiently long relative to the wavelength of the  
 4518  $TE_{01}$  modes than perfect reflection can be achieved. This derivation is quite lengthy  
 4519 and complex and is presented in full in [77]. Here, we shall simply explain the resulting

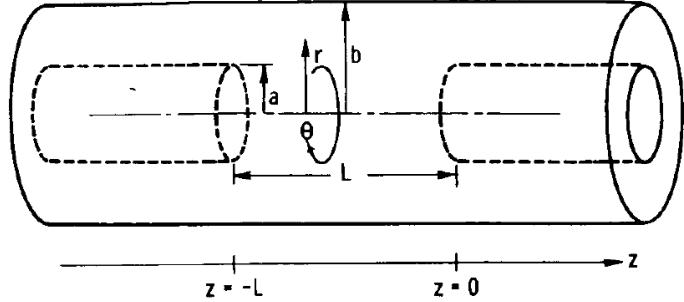


Figure 6.10: The simplified geometry of an open cavity with coaxial terminations. Figure from [77].

4520 conditions on the partition radius for perfect reflection.

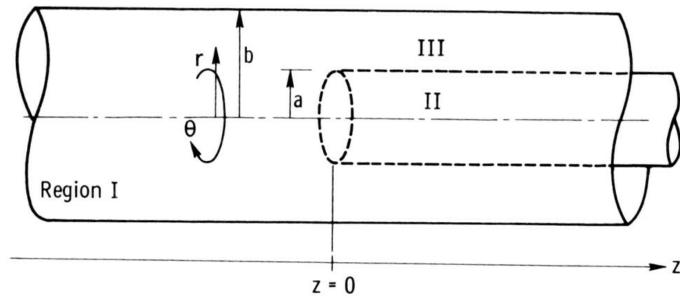


Figure 6.11: Electric field regions for the open cavity boundary value problem. Figure from [77].

4521 The open cavity boundary value problem is solved by expressing the forms of the  
 4522 electric fields in the different regions of the cavity and requiring that the electric fields are  
 4523 continuous. There are effectively three distinct regions in the open cavity corresponding  
 4524 to the central cavity volume, the inner coaxial volume, and the outer coaxial volume (see  
 4525 Figure 6.11).

4526 In Region I, the boundary conditions are those of a cylindrical waveguide, and we  
 4527 require that  $E_\phi$  for the  $TE_{0m}$  modes go to zero at the cavity wall ( $r = b$ ). This requires  
 4528 that  $J'_{0m}(k_{c0m} b) = 0$ . We aim to solve for the radius  $a$  in the specific situation where the  
 4529  $TE_{01}$  mode can propagate but all other  $TE_{0m}$  modes are below the cutoff frequency for  
 4530 the circular waveguide. This is equivalent to requiring

$$3.832 < k_{c0m} b < 7.016, \quad (6.41)$$

4531 where the numbers 3.832 and 7.016 correspond to the first and second zeros of the Bessel

4532 function (see Table 6.1).

4533 In Region II the boundary conditions are those of a cylindrical waveguide, but with  
4534 a smaller radius. The condition that  $E_\phi = 0$  at the cylindrical partition radius is that  
4535  $J'_{0m}(k_{c0m}a) = 0$ . To ensure perfect reflection, we want all modes in Region 1 of the cavity  
4536 to be below the cutoff frequency of the circular waveguide formed by the inner volume of  
4537 the coaxial terminator. Therefore, we consider the solutions where

$$k_{c0m}a < 3.832. \quad (6.42)$$

4538 Finally, in Region III the boundary condition are those of a coaxial waveguide. We  
4539 need to guarantee that  $E_\phi = 0$  at both  $r = b$  and  $r = a$ , which involves finding the  
4540 eigenvalues of the following equation

$$J'_0(k_{c0m}a)Y'_0(k_{c0m}b) - J'_0(k_{c0m}b)Y'_0(k_{c0m}a) = 0, \quad (6.43)$$

4541 where  $Y'_0$  the zeroth-order derivatives of the Bessel function of the second kind. The  
4542 solutions to this equation depend on the value of the ratio  $b/a$ . The approximate solution  
4543 is given by

$$\delta_n a \simeq \frac{n\pi}{b/a - 1}, \quad (6.44)$$

4544 where  $\delta_n$  are eigenvalues of Equation 6.43. Similar to Region II, we are interested in  
4545 solutions for which the TE<sub>01</sub> modes of Region I are below the cutoff of Region III.  
4546 Therefore, we require that

$$k_{c0m} < \delta_1. \quad (6.45)$$

4547 In general, one has some freedom in specifying the value of  $b/a$ . A value typically used  
4548 in practice is  $b/a = 2.082$ , which corresponds to positioning the radius of the cylindrical  
4549 partition at the maxima of the TE<sub>01</sub> electrical fields.

4550 Using the constraints from the three field regions one can develop a coaxial terminator  
4551 that acts as a virtual perfectly conducting surface for the TE<sub>01</sub> modes. The only required  
4552 inputs are the desired frequency of the TE<sub>011</sub> mode and a choice for the value of  $b/a$ .

#### 4553 **6.4.2 Mode Filtering**

4554 The general case of an electron coupling to a resonant cavity is complicated. This is  
4555 because cavities contain an infinite number of resonant modes, which for higher order  
4556 modes, have couplings to the electron with a complex spatial dependence. The danger is

4557 that improper modeling of the electron's coupling to the cavity can lead to systematic  
4558 errors in the CRES measurements that prevent a high-resolution measurement of the  
4559 electron's kinetic energy. This in part drives the preference for a single-mode cavity  
4560 experiment that uses only the electron's coupling to the TE<sub>011</sub> mode to perform CRES,  
4561 assuming that sufficient information on the electron's position can be obtained with a  
4562 single mode.

4563 The TE<sub>011</sub> mode is in a region where there are relatively few other modes to which  
4564 the electron could couple(see Figure 6.8). However, one can see that the frequency of  
4565 the TE<sub>011</sub> is perfectly degenerate with the TM<sub>111</sub> mode, which means that electrons will  
4566 inevitably couple to both modes if they have the correct cyclotron frequency.

4567 The magnitude of the impact of the electron coupling to both TE<sub>011</sub> and TM<sub>111</sub> is  
4568 currently unknown. To first order an electron coupling to more both modes will lose more  
4569 energy overtime, which can be measured by observing the frequency chirp rate of the  
4570 signal. This effect may be small enough to be negligible or simple enough to model that  
4571 the cavity can be treated as an effective single-mode cavity. Alternatively, the one could  
4572 consider devising a coupling scheme that is sensitive to both the TE<sub>011</sub> and the TM<sub>111</sub>  
4573 modes. By measuring the coupling of the electron to both modes more information on  
4574 the position of the electron could be obtained, which could improve the position and  
4575 energy resolution of the CRES measurements.

4576 A different approach is the mode filtering approach, which seeks to obtain a single  
4577 TE<sub>011</sub> mode cavity using perturbations to the cavity walls that selectively impede the  
4578 TM modes, while leaving the TE modes mostly unperturbed. The type of perturbations  
4579 required can be determined by visualizing the surface currents induced in the cavity  
4580 walls by each type of mode (see Figure 6.12). By definition, all TM have electric fields  
4581 directed along the vertical axis of the cylindrical cavity, which means that perturbations  
4582 that impede currents in this direction will modify TM resonances. On the other hand,  
4583 the TE<sub>01</sub> modes induce azimuthal currents in the cavity walls, therefore, it is possible to  
4584 break the degeneracy between TE<sub>01</sub> and TM<sub>11</sub> using a cavity perturbation that impedes  
4585 axial currents, but does not affect the flow of azimuthal currents.

4586 Figure 6.12 shows two cavity design concepts that achieve this selective current  
4587 perturbation. The resistive approach inserts a series of thin dielectric rings into the walls  
4588 of the cavity that introduces a resistive and capacitive impedance to the longitudinal  
4589 currents, while leaving azimuthal current paths intact. Cavities of this type with high  
4590 TE<sub>01</sub> Q's have also been constructed by tightly wrapping a thin, dielectric coated wire  
4591 around a mold to form the cavity wall. An alternative method is to introduce an inductive

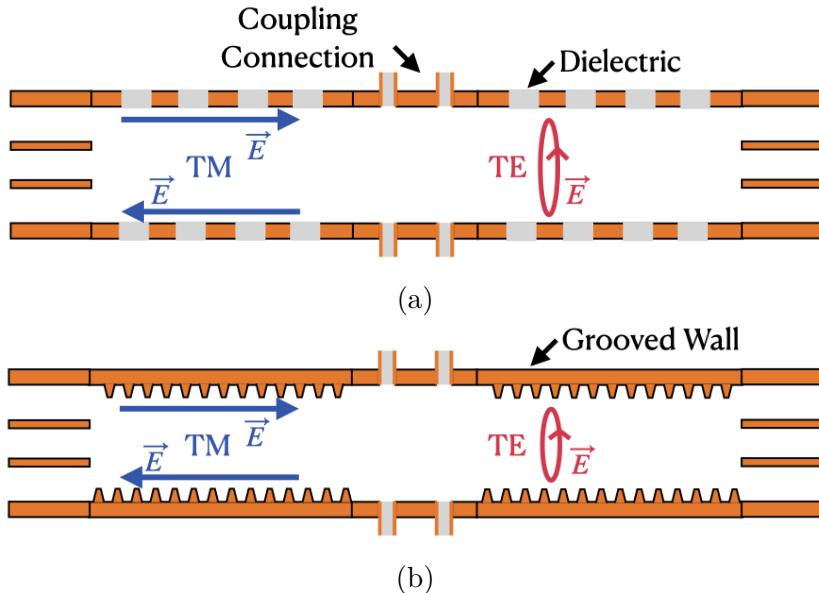


Figure 6.12: Two mode filtering concepts to break the degeneracy of  $TE_{01}$  and  $TM_{11}$  modes. The resistive approach uses dielectric materials to impede currents that travel vertically along the cavity while leaving azimuthal currents unperturbed. An alternative approach is to impede the currents using grooves cut into the cavity wall, which achieve the same effect with an inductive impedance.

4592 impedance by cutting grooves or a thread pattern on the inside wall of the cavity. For  
 4593 reasons of manufacturability and compatibility with tritium the grooved cavity approach  
 4594 is the preferred method for mode-filtered cavity construction by Project 8.

#### 4595 6.4.3 Simulations of Open, Mode-filtered Cavities

4596 A candidate design for a single  $TE_{011}$  mode CRES experiment is a cavity that utilizes  
 4597 the coaxial terminations combined with a mode-filtering wall. The first step towards  
 4598 validating that a cavity that combines these two design features will operate as expected  
 4599 is a thorough simulation effort for which finite element method (FEM) simulation software  
 4600 is invaluable. The primary tool for electromagnetic FEM calculations inside Project 8 is  
 4601 Ansys HFSS, which has a robust and well-established eigenmode solver that can identify  
 4602 the resonant frequencies and associated Q-factors for given structure.

4603 Four variations of a cavity design with a  $\sim 1$  GHz  $TE_{011}$  resonance were implemented  
 4604 in HFSS (see Figure 6.13). The four designs include a standard cylindrical cavity, an  
 4605 open cavity with smooth walls, an open cavity with resistive walls, and an open cavity  
 4606 with grooved walls. The relevant design parameters are summarized in Table 6.3. All

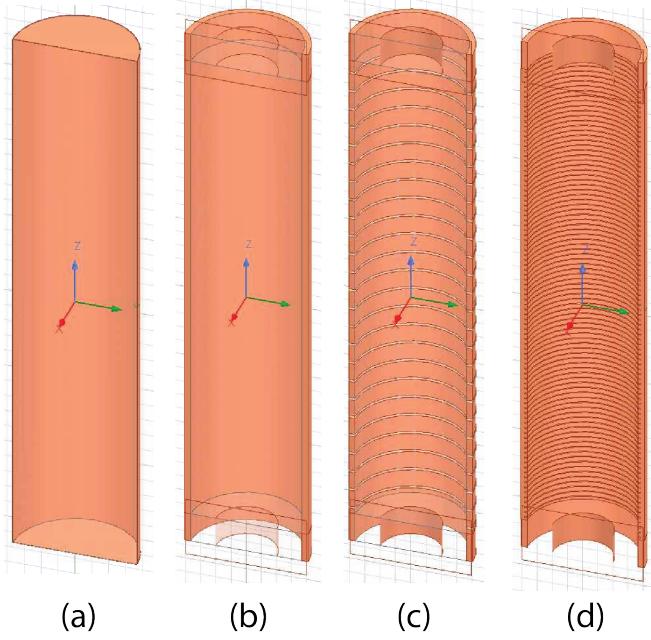


Figure 6.13: Four cavity design variations. (a) is a standard sealed cylindrical cavity, (b) is an open cavity with smooth walls, (c) is an open cavity with resistive walls, and (d) is an open cavity with grooved walls. The main cavity and coaxial terminator parameter are identical for all four cavities.

<sup>4607</sup> cavities were simulated using copper walls and filled with a vacuum dielectric. The  
<sup>4608</sup> identities of the resonant modes found by HFSS were validated by visual inspection of  
<sup>4609</sup> the electric and magnetic field patterns and by comparison to analytical calculations of  
<sup>4610</sup> the mode frequencies.

Table 6.3: A table of cavity design parameters used for HFSS simulations.

Name	Qty.	Unit	Description
$D_{\text{cav}}$	326.4	mm	Cavity diameter
$L_{\text{cav}}$	1668.0	mm	Cavity length
$D_{\text{term}}$	200.2	mm	Inner diameter of coaxial terminator
$L_{\text{term}}$	100.0	mm	Terminator length
$l_{\text{die}}$	8.3	mm	Dielectric spacer thickness
$\Delta l_{\text{die}}$	66.7	mm	Distance between dielectric spacers
$l_{\text{groove}}$	3.0	mm	Groove height
$d_{\text{groove}}$	9.0	mm	Groove depth
$\Delta l_{\text{groove}}$	18.3	mm	Distance between grooves

<sup>4611</sup> The results of the HFSS simulations validate our predictions of the resonant behavior  
<sup>4612</sup> of an open, mode-filtered cavity developed in the preceding sections (see Figure 6.14) One

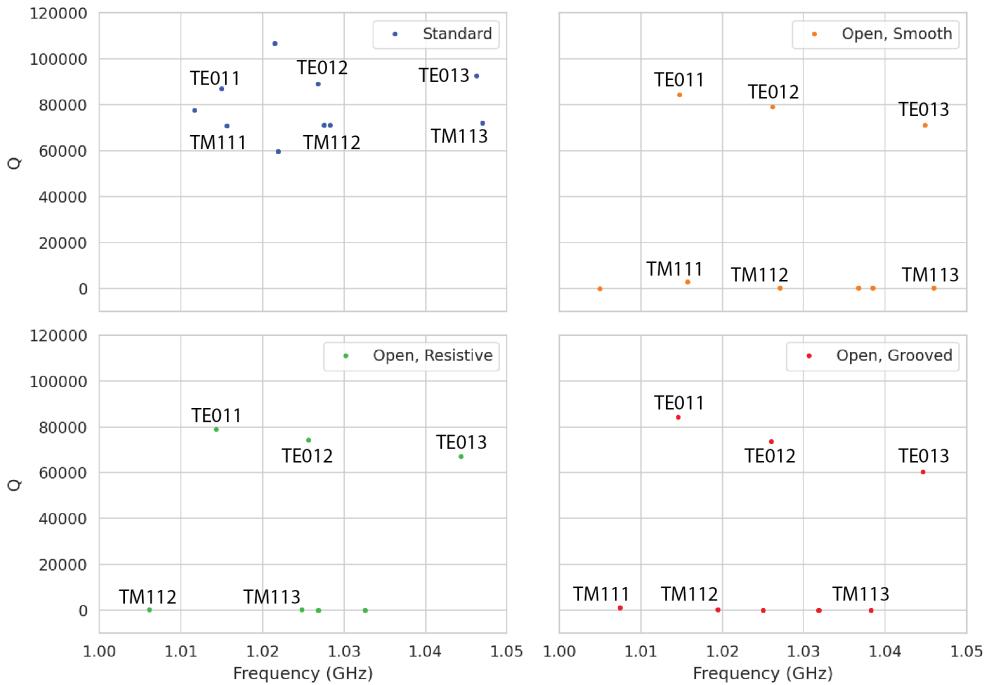


Figure 6.14: The frequencies and Q-factors of the resonant modes identified by HFSS for the cavity variations shown in Figure 6.13. The fully-sealed cavity with smooth walls has several high-Q modes near the  $TE_{01}$  resonance. Introducing the open-termination preserves the Q-factors of the  $TE_{01\ell}$  modes and suppresses the Q-factors of the modes whose boundary conditions do not match the cylindrical partition. Both the resistive and grooved wall perturbations shift the resonant frequencies of the TM modes away from the  $TE_{01}$  mode. By properly tuning the geometry of the grooves or the resistive spacers several MHz of frequency separation can be achieved.

4613 can see that for a standard cavity the  $TE_{01}$  and the  $TM_{11}$  are degenerate in frequency  
 4614 with relatively high Q-factors. The open-ended cavity preserves the high Q-factors of  
 4615 the  $TE_{01}$  modes, while the other modes, since their boundary conditions do not match  
 4616 the coaxial geometry, have their Q-factors suppressed. One can see that the effect of  
 4617 the resistive and inductive mode-filtering schemes is to effectively shift the resonant  
 4618 frequencies of the  $TM_{11}$  modes below those of the associated  $TE_{01}$  modes, which breaks  
 4619 the degeneracy. Optimization of the dielectric spacer or groove parameters can ensure  
 4620 that the  $TE_{01}$  mode is isolated from other modes by  $O(10)$  MHz, which provides sufficient  
 4621 bandwidth for a measurement of the tritium spectrum endpoint.

4622 Further optimization of the cavity design requires a more detailed cavity simulation  
 4623 that includes the cavity coupling mechanism as well as other geometry modifications  
 4624 required for integration into the magnetic and tritium gas subsystems. Perhaps more

4625 important is the development of the capability to simulate the interaction of electrons  
4626 with the cavity so that simulated CRES signals can be generated using cavities designed  
4627 for CRES measurements. Simulated CRES signals can then be used to estimate the  
4628 neutrino mass sensitivity of the experiment, which allows for the optimization of the cavity  
4629 design towards the configuration that provides the best measurement of the neutrino  
4630 mass.

## 4631 **6.5 Single-mode Resonant Cavity Measurements**

4632 Measurement test stands play an important role in the research and development process  
4633 that cannot be replaced by simulations. For example, constructing a prototype CRES  
4634 cavity forces one to consider important practical issues such as manufacturability and  
4635 machine tolerances that may require modifications to the design. Furthermore, by  
4636 comparing laboratory measurements of a real cavity to simulations, one can quantify  
4637 the impact of imperfections and real-life measurement systematics, which allows for  
4638 more accurate sensitivity estimates of the experiment. Lastly, the development of these  
4639 prototypes helps to build the necessary experience and expertise within the collaboration  
4640 required for more complicated experiments to succeed.

4641 In this spirit a prototype cavity was constructed to demonstrate the open, mode-  
4642 filtered cavity concept explored in the previous sections. The primary goal of the  
4643 measurements was to validate that an open, mode-filtered cavity suppressed the  $\text{TM}_{11}$   
4644 modes as predicted by HFSS simulations.

### 4645 **6.5.1 Cavities and Setup**

4646 Two rudimentary, cavities were constructed using segments of copper pipe available from  
4647 McMaster-Carr (see Figure 6.15). The design consists of copper pipes of two diameters.  
4648 The larger diameter pipe forms the main cavity wall and the smaller diameter pipe is  
4649 used to create a coaxial termination. The diameter of the outer pipe was chosen to  
4650 produce a  $\text{TE}_{011}$  resonance of approximately 6 GHz, while the diameter of the smaller  
4651 pipe was selected based on the open termination criteria introduced in Section 6.4.1. The  
4652 approximate diameters and lengths of the copper pipe are summarized in Table 6.4.

4653 Coupling to the cavity was achieved using a hand-formable segment of coaxial cable  
4654 stripped at one end to form a loop antenna. This was inserted into a small hole located  
4655 at the center of the main cavity wall. The coaxial terminators were supported inside the

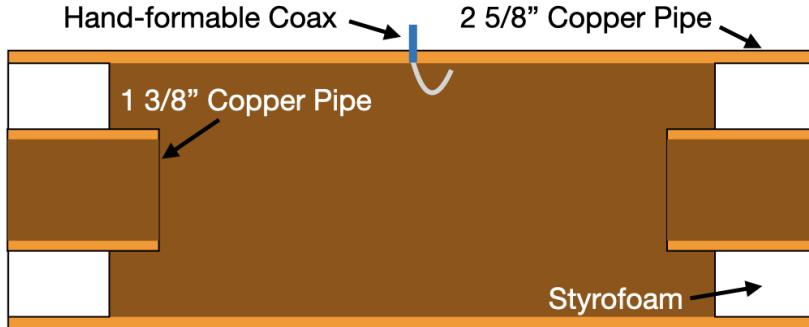


Figure 6.15: A cartoon depicting the design of the open-ended cavity prototype designed to operate at approximately 6 GHz. The main cavity wall was composed of a single copper pipe. A mode-filtered version of this cavity was constructed by

<sup>4656</sup> main cavity by carving a spacer from polystyrene foam (styrofoam) so that they could  
<sup>4657</sup> be easily inserted into the cavity and repositioned. The dielectric constant of styrofoam  
<sup>4658</sup> is quite close to air at microwave frequencies so this is expected to have minimal impact  
<sup>4659</sup> on the resonant properties of the cavity.

Table 6.4: A table of parameters describing the cavity prototypes. Certain values such as the cavity length and the distance between dielectric spacers are approximate due to variation in the machining of the copper. In particular, the filtered cavity was constructed from conducting copper segments that varied in size from 1.50" to 1.85".

Name	Qty.	Unit	Description
$D_{\text{cav}}$	2.625	in	Cavity diameter
$L_{\text{cav}}$	$\approx 13$	in	Cavity length
$D_{\text{term}}$	1.375	in	Inner diameter of coaxial terminator
$L_{\text{term}}$	1.575	in	Terminator length
$l_{\text{die}}$	0.75	in	Dielectric spacer thickness
$\Delta l_{\text{die}}$	$\approx 1.50$ to $1.85$	in	Distance between dielectric spacers

<sup>4660</sup> The actual length of the cavity is given by the distance between the inner edges of the  
<sup>4661</sup> coaxial terminations. The length of the outer section of pipe that forms the main wall of  
<sup>4662</sup> the cavity is approximately 16" in length which leads to a cavity length of  $\approx 13"$  when  
<sup>4663</sup> both terminators are inserted in the cavity. Because the terminators were not rigidly  
<sup>4664</sup> mounted this distance is only approximate, however, the uncertain length of the cavity  
<sup>4665</sup> will not prevent us from validating the open cavity design.

<sup>4666</sup> Along with the smooth-walled open cavity a resistively mode-filtered cavity was  
<sup>4667</sup> constructed by creating dielectric spacers out of segments of clear PVC pipe (see Figure  
<sup>4668</sup> 6.16). The spacers were machined such that the conductive segments of the cavity would

4669 be separated by 0.75" when the cavity was fully assembled. Due to variations in the  
 4670 lengths of the copper segments that make up the cavity wall the distance between spacers  
 4671 has significant variation with average value of about 1.7". Eight total spacers were used  
 4672 to build the cavity, which when assembled was approximately 16" in total length similar  
 to the non-filtered cavity.

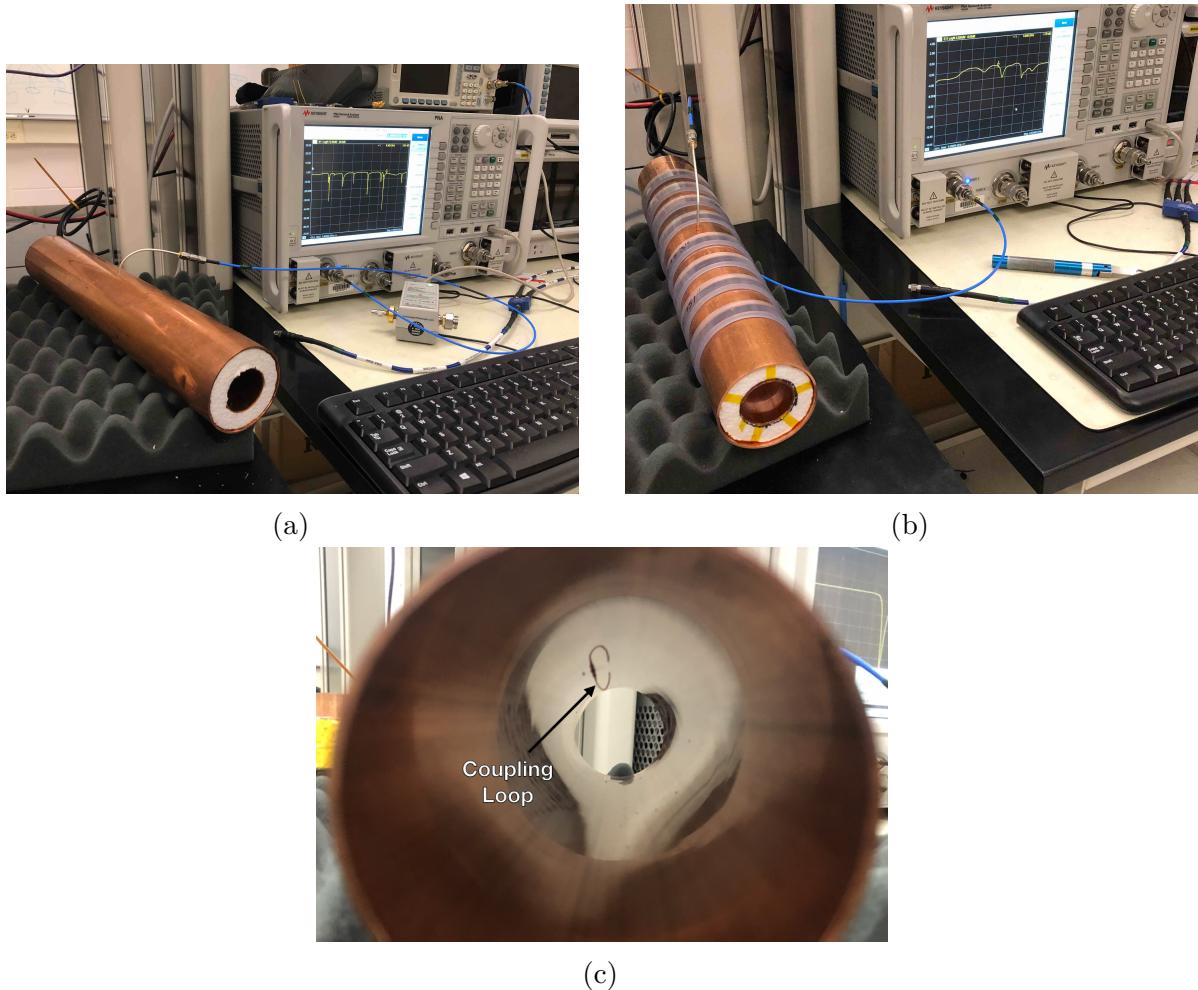


Figure 6.16: Images depicting the measurement of the filtered and non-filtered open cavities using the VNA. The coupling loop in the figure is shown in the TE orientation.

4673 Measurements of both cavities were performed using a VNA connected to the cavity  
 4674 coupling probe (see Figure 6.16). By measuring the return loss over a range of frequencies  
 4675 one can measure the frequencies and relative Q-factors of the resonant modes in the  
 4676 cavity. Due to the opposite polarity of the electric fields for the TE and TM modes,  
 4677 the loop coupling probe must be rotated 90° to change the polarity of the loop antenna.  
 4678 When the antenna is oriented such that the loop opening faces the ends of the cavity, it  
 4679

4680 couples primarily to the TE modes which have magnetic fields directed along the long  
 4681 axis of the cavity (see Figure 6.16). If the coupling loop is turned by  $90^\circ$  from where  
 4682 it is shown in the image then it will couple to the TM modes which have azimuthally  
 4683 directed magnetic fields. In this way both the TE and TM resonances can be measured  
 4684 independently.

### 4685 **6.5.2 Results and Discussion**

4686 The primary analysis method for the prototype cavities involved simply visualizing the  
 4687 return loss measured by the VNA and comparing between the filtered and non-filtered  
 4688 cavities. Since the resonances measured by the VNA are not labeled, there is some  
 4689 uncertainty about the true identities of the modes measured by the VNA. To help with  
 4690 this we performed a simulation of the simplest possible cavity that could be created from  
 4691 the prototype components, which is a fully open cavity created by simply removing the  
 4692 coaxial inserts from the non-filtered cavity configuration. The fully open cavity with the  
 4693 as-built dimensions was simulated in HFSS to get estimates on the positions of the  $\text{TE}_{011}$   
 4694 and  $\text{TM}_{111}$  modes (see Figure 6.17).

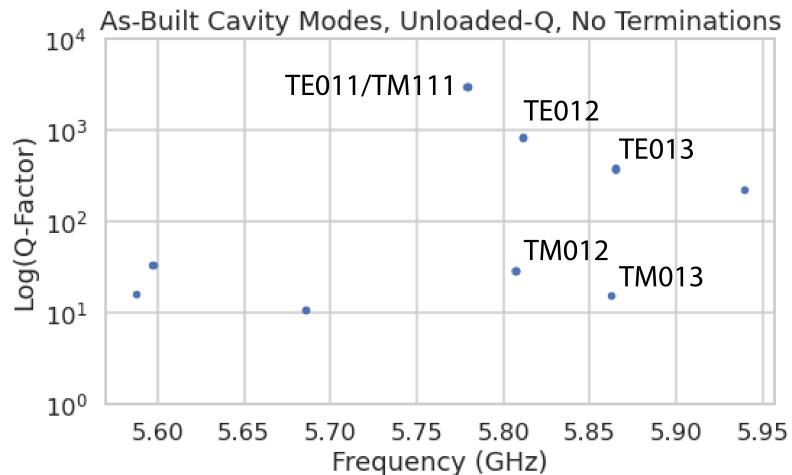


Figure 6.17: HFSS simulation results for a the as-built cavity with the coaxial terminators removed. The  $\text{TE}_{011}/\text{TM}_{111}$  frequency is approximately 5.78 GHz.

4695 Simulation of the fully open cavity shows that the  $\text{TE}_{011}/\text{TM}_{111}$  modes have a  
 4696 frequency of approximately 5.78 GHz in the fully open cavity. If the frequency of this  
 4697 mode is compared to the measurments of the fitered and non-filtered cavities with the  
 4698 terminators removed we can easily identify the  $\text{TE}_{011}$  mode at approximately 5.75 GHz

4699 (see Figure 6.18).

4700 For the non-filtered cavity one sees that the  $TE_{011}$  mode is degenerate in frequency  
 4701 with what appears to be a doublet of TM modes located at the  $TM_{111}$  frequency position.  
 4702 This doublet is actually the  $TM_{111}$  mode, which has two polarizations with opposite  
 4703 polarizations. Because the pipe used to construct the cavity is not perfectly round, the  
 4704 frequency degeneracy between the two polarizations is broken producing the doublet  
 peak. In the case of the filtered cavity with no terminators there is an isolated TE

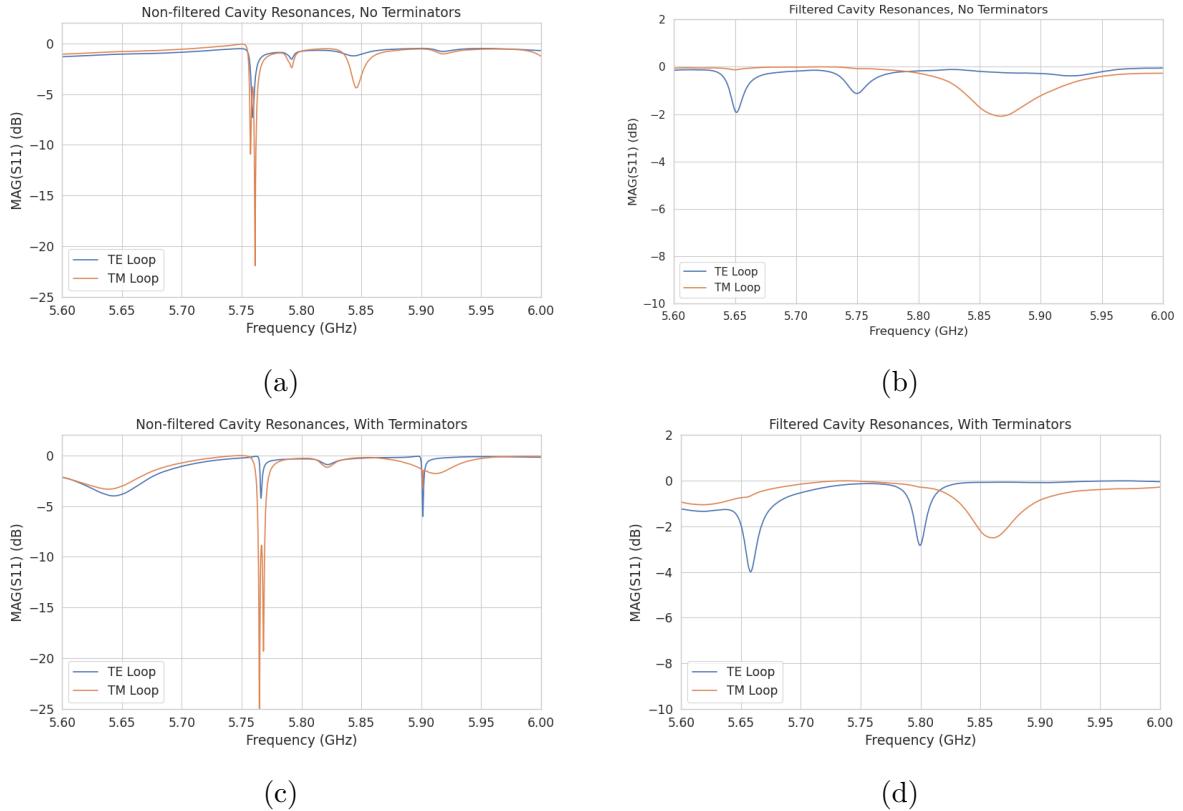


Figure 6.18: Measurements of the filtered and non-filtered prototype cavities acquired with the VNA.

4705  
 4706 resonance at 5.75 GHz that appears to be the  $TE_{011}$ , however, there is no apparent  $TM_{111}$   
 4707 doublet at the same frequency. This is what one would expect if the mode-filtering was  
 4708 effective at suppressing the TM modes. There is a notable difference in the Q of the  
 4709  $TE_{011}$  resonance for the non-filtered and filtered cavities indicated by the relative widths  
 4710 of the resonances. This is likely caused by the large width of the dielectric spacers that  
 4711 are partially impeding the TE modes. When the terminators are inserted into the cavity  
 4712 one sees that Q-factors of the modes improves as expected, by noticing the narrowing of  
 4713 the peaks compared to the no terminator plots.

<sup>4714</sup> In conclusion, one see from these cavity measurements that, in principle, mode-  
<sup>4715</sup> filtering can be used to separate the TE<sub>011</sub> resonance from the degenerate TM<sub>111</sub> mode in  
<sup>4716</sup> combination with the an open cavity design. The ideal next step would be to construct a  
<sup>4717</sup> open, mode-filtered cavity that could be used to perform CRES measurements. In order  
<sup>4718</sup> to study the coupling of an electron to the isolated TE<sub>011</sub> mode.

4719 **Chapter 7 |**

4720 **Conclusion and Future Prospects**

4721 In this dissertation we have discussed research and development efforts towards the  
4722 development of a scalable CRES measurement technology that can be used to build a  
4723 CRES experiment at cubic-meter scales with sensitivity to neutrino masses of 40 meV.  
4724 The primary contributions of my dissertation are the development and analysis of signal  
4725 reconstruction algorithms for an antenna array based CRES experiment [79], which leads  
4726 to estimates of the neutrino mass sensitivity; the development of a synthetic cyclotron  
4727 radiation antenna (SYNCA) [44], which allowed for laboratory validation of antenna  
4728 array CRES simulation models [7]; and the development of an open-ended cavity design  
4729 compatible with atomic tritium for a cavity based CRES experiment. A measurable  
4730 impact of this work is the transition of the Project 8 collaboration’s experimental plan  
4731 from an antenna array based approach to a cavity based approach, where my work played  
4732 a key role in demonstrating the significantly higher cost and complexity of the antenna  
4733 array experiment.

4734 The transition from antenna arrays to cavities requires a new set of demonstrator  
4735 experiments to make incremental progress towards a 40 meV measurement of the neutrino  
4736 mass. At the time of writing, the near-term plan of Project 8 is to design and construct a  
4737 small-scale cavity CRES experiment utilizing the 1 T magnet installed in the UW-Seattle.  
4738 This cavity is designed to have a TE011 resonance with a frequency of about 26 GHz with  
4739 a length-to-diameter ratio that mimics the larger cavities intended for the pilot-scale and  
4740 Phase IV experiments. The goal of this experiment is to demonstrate cavity CRES as  
4741 well as validate models of CRES systematics using electrons from  $^{83m}\text{Kr}$  and an electron  
4742 gun. Though the primary goal is demonstration, near-term physics measurements are  
4743 available in the form of high-resolution measurements of the  $^{83m}\text{Kr}$  conversion spectrum  
4744 of interest to the KATRIN collaboration.

4745 Furthermore, Project 8 is currently constructing a low-frequency CRES setup located  
4746 at Yale University to better understand the principles of cavity based CRES at lower

4747 magnetic fields. The Low, UHF Cavity Krypton Experiment at Yale (LUCKEY) is  
4748 a 1.5 GHz cavity CRES experiment the will use conversion electrons from  $^{83m}\text{Kr}$  to  
4749 perform CRES measurements at the lowest frequencies ever attempted with the technique.  
4750 LUCKEY will validate frequency scaling models developed by Project 8 and will pave  
4751 the way for the future Low-Frequency Apparatus (LFA), which will be a larger, 1 GHz  
4752 cavity CRES experiment that includes a molecular tritium source. The target for the  
4753 LFA is a measurement of the neutrino mass with a sensitivity of approximately 0.2 eV,  
4754 which will build towards the atomic pilot-scale CRES experiment.

4755 In parallel to the development of cavity CRES is the development of the atomic  
4756 tritium source. Recent demonstrations of the production of atomic hydrogen are excellent  
4757 steps towards the atomic tritium production needed for the pilot-scale experiment. One  
4758 area of future study includes the development of a more detailed unstanding of the  
4759 efficiency of atomic hydrogen production. Near-term plans include the development of a  
4760 magnetic, evaporatively cooled beamline, as well as the prototyping of a Halbach array  
4761 atoms trap. Nearly all of the components of the atomic tritium system will require  
4762 demonstration before the complete system can be built. The long-term goal of the  
4763 atomic tritium work is to construct a full atomic tritium prototype that demonstrates  
4764 the production, cooling, trapping, and recycling of tritium at the rates needed for the  
4765 pilot-scale experiment.

4766 More broadly, the long-term goal of the Project 8 collaboration is to fully develop  
4767 both the atomic tritium and cavity CRES technologies so that both can be combined in  
4768 a pilot-scale CRES experiment. It is envisioned that this process will take approximately  
4769 10 years for both atomic tritium and cavity CRES. After these developments comes  
4770 the pilot-scale experiment which will be the first CRES experiment that simultaneously  
4771 demonstrates all the required technologies for Phase IV. Scaling to Phase IV with cavity  
4772 CRES will require the construction of multiple copies (approximately 10) of the pilot-scale  
4773 experiment to obtain sufficient statistics for 40 meV sensitivity.

4774 Development of the CRES experimental technique by Project 8 has led to new  
4775 experiments utilizing the CRES technique for basic physics research, such as the  $^6\text{He}$ -  
4776 CRES collaboration [80], and has also found applications as a new approach to x-ray  
4777 spectroscopy [81]. Recently, a new experimental effort called CRESDA has begun in  
4778 the UK to develop new quantum technologies applied to CRES measurements for the  
4779 neutrino mass [82]. This flourishing of new experimental efforts based on the CRES  
4780 technique is likely to continue as Project 8 continues to develop the technique towards  
4781 its neutrino mass measurement goal.

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