REACTIVE RF TUNING FOR COMPENSATION OF A DETUNED ACCELERATING CAVITY*

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Abstract

The resonant frequency of an accelerating RF cavity is detuned from the desired frequency by certain physical disturbances, such as thermal and other mechanical wall distortions. Cavity wall distortions due to *microphonics* (acoustic vibrations) and the *Lorentz force* (radiation pressure) can be serious problems in pulsed RF operation of superconducting (SRF) cavities with thin cavity walls and a high quality factor. The resulting detuning results a change of input reactance. The offset reactance at the cavity input may be tuned out properly with a reactive element in the input transmission line, so that the generator RF power can be delivered efficiently to the cavity. A fast response electrical tuner may be built for compensating high frequency detuning without any mechanical coupling. We present our concept and design, as well as discussions on reactive compensation through the cavity input transmission line in the context of other work.

1 INTRODUCTION

Accelerator performance depends on precise control of the amplitude and phase of the accelerating cavity RF fields. An optimally beam loaded cavity can still be detuned, resulting in mismatched coupler input, due to various mechanical forces. This mismatch needs to be corrected in real time in order to ensure the efficient operation of the accelerator in terms of the electrical energy and the beam quality. This kind of compensation is of particular interest in superconducting cavities with very high quality factors, experiencing distortions from both the *Lorentz force* and *microphonics*. Piezo-mechanical tuners in development for fast compensation of detuned superconducting cavities [1], however, drive additional acoustic vibrations. Purely electrical tuners may avoid parasitic acoustic noise, and the compensation may be applied outside of the cryomodule, relaxing maintenance difficulties.

There have been investigations of the tuning of beam loaded cavities aimed at placing the tuning elements in the input RF transmission line outside the vacuum, such as that of Krejcik[2]. The solution Krejcik presents for static compensation for normal conducting cavities is particularly interesting as a supplement for real-time control of SRF cavities. For superconducting cavities, a major advantage remains the practical aspect of not requiring access to the cryomodule, which is usually difficult once it is mechanically sealed.

In this paper, non-mechanical frequency compensation of a detuned cavity with a single element in the transmission line is presented: it uses an adjustable shunt reactive element, a capacitor or an inductor, in the feed transmission line.

2 SINGLE ELEMENT TUNING OF DETUNED CAVITY

If a cavity is detuned by mechanical strain in axial direction, mainly C or L in its equivalent LC resonant circuit changes. A capacitance change of ΔC results detuning Δf. This is illustrated on a Smith chart in Figure 1. Detuned cavity impedances for cases $\Delta f > 0$, and $\Delta f < 0$ are Zc^+ and Zc^-, respectively.

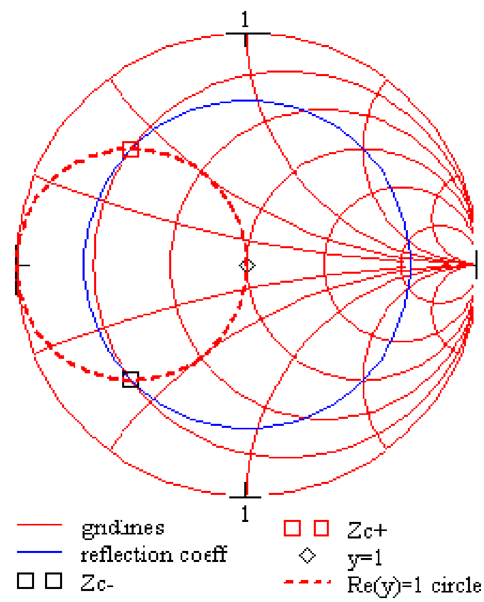

Figure 1 – Smith chart representation of the reactive tuning shown for all cases

The cavity input impedances are also on the $Re(y)=1$ circle and repeats if the distance from the cavity is an integer multiple of the transmission line half wavelength $\lambda_g/2$. The input impedance Zc^+ (or Zc^-) can be matched with a shunt susceptance or a series reactance at d when it is moved to $y=1$ circle or $z=1$ circle, respectively. This is done if the distance satisfies the following for $N = 1, 2, \ldots$:

$d=N\lambda_g/2$ for shunt element tuning (1a)

$d=(1/2+N)\lambda_g/2$ for the series element tuning (1b)

by transforming the Y_c $(=1/Z_c)$ to $y=1$ or the Z_c to $z=1$, at the chart center, respectively. Figure 2 shows the proposed single element reactive compensation of the detuned cavity with the equivalent circuit of the resonant cavity. The shunt element tuning is shown in (a)(b)(c) and the series element tuning is shown in (d)(e)(f).

For frequency tuning by a series element, the distance d from the cavity input that satisfies the relationship [3]

$$R_c = \mathrm{Re}\left[\frac{Z_o(Z_c + Z_o \tanh(\gamma d))}{Z_o + Z_c \tanh(\gamma d)}\right] \tag{2}$$

where Z_c is the cavity input impedance and R_c is the source impedance or the transmission line characteristic impedance. The propagation constant $\gamma = \alpha + j\beta$ where α is the attenuation constant and β is the phase constant. A similar process may be used for shunt element tuning with admittances. Two solutions exist on d for the matching in Eq. (2). However, only one solution provides a constant distance for compensation of any detuning. The other solution d is not a constant and changes for different detuning. If $\Delta f > 0$ ($\Delta f < 0$), the tuning elements are $jB>0$ ($jB<0$) for the shunt element and $jX>0$ ($jX<0$) for the series element, respectively.

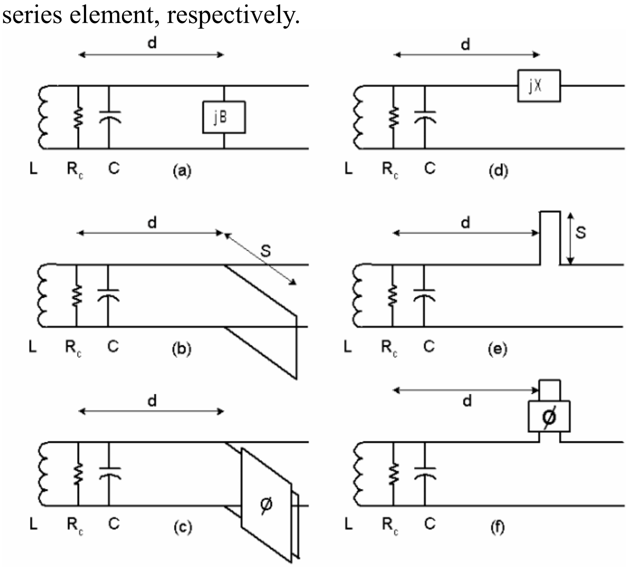

Figure 2: Reactive compensation of a detuned resonator. Realization of (a)(b)(c) shunt element tuning and (d)(e)(f) series element tuning. For shunt tuning (series tuning), the shunt stub (b) (series stub (e)) is substituted for the susceptance (a) (reactance (d)) in Figure 2. For each case, a stub can be substituted by a short circuited phase shifter (c)(f).

The phase shifter phase delay in the shunt stub can be made identical to the phase shift of the tuning element susceptance. The short circuited phase shifter may provide a phase shift $\pm\Delta\phi$ that covers the required susceptance $\pm B$ where $\Delta\phi$ is the half of the phase shifter shifting range..

The cavity input admittance seen at the input coupler with n:1 transformer ratio is given as

$$\frac{Y_c'}{Y_o} = \frac{n^2 G_c}{Y_o} + \frac{jn^2}{Y_o}\sqrt{\frac{C}{L}}\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right) \tag{3}$$

where

$$\omega_o = \frac{1}{\sqrt{LC}}$$

and the transmission line characteristic admittance is

$$Y_o = \beta_c n^2 G_c$$

where β_c is the coupling constant of the cavity and $\beta_c = 1$ has been assumed in present consideration. Now, for the shunt tuning with $d = N\lambda_g/2$, the shunt susceptance jB is equal to the cavity input susceptance jB_c to maintain a tuned shunt resistance of the cavity. Since the short stub input impedance for a lossless case $Y_{in} = -jY_o \cot(\beta s)$, the above relationship is expressed as,

$$\cot(\beta s) = \frac{-n^2}{Y_o}\sqrt{\frac{C}{L}}\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right) \tag{4}$$

The phase term βs is the short circuited shunt stub phase delay that can make the input admittance equivalent to the susceptive loading that tunes out jB at the cavity input. Similar analysis can be made for the series tuning.

3 TRANSMISSION LINE

The transmission line between the cavity and the tuning stub supports a standing wave due to the multiple backward and forward reflected waves at the cavity and at the T-junction, respectively. In real RF systems, the waveguide length can be many times the wavelength at the operating frequency, but the resulting time delay is negligible compared to the intrinsic cavity fill time.

For an RF pulse length that is much greater than the transmission line length, the incident voltage to the T-junction will be partially reflected back to the source and partially transmitted to the cavity through the transmission line. The standing wave voltage and the voltage delivered to the cavity through the waveguide at a point $z=z'$ can be expressed in terms of the forward propagating waves in the transmission line as

$$V_w(t) = \sum_i^{\infty}\left(\Gamma_1\Gamma_2 V_i^f\left(t - \frac{z'}{v} - \frac{d}{v}\right) + \Gamma_2 V_{i+1}^f\left(t - \frac{z'}{v} - \frac{d}{v}\right)\right)$$

$$V_c(t) = \sum_i^{\infty} T_2 V_i^f(t - d/v)$$

where Γ_1 is the voltage reflection coefficient at the T-junction with the shunt element and the transmission coefficient $T_2 = 1 + \Gamma_2$ is the voltage transmission coefficient at the cavity input where Γ_2 is the reflection coefficient at the cavity input. The i-th forward wave is

$$V_i^f(t) = \Gamma_1 V_i^r(t + \frac{z'}{v} - \frac{d}{v})$$

where for $i=1$, $V^f{}_1 = T_1 V_o e^{j\omega t}$ and *the i-th* reflected wave

$$V_i^r(t) = \Gamma_2 V_{i-1}^f(t - \frac{z'}{v} - \frac{d}{v})$$

The standing wave in the transmission line and the power transferred to the cavity are shown in Figure 3. For both cases, the capacitive compensation is made for a frequency detuning of $\Delta f = -f/Q$. The time delay on the power transfer may be negligible in long pulse applications even in a fairly long transmission lines.

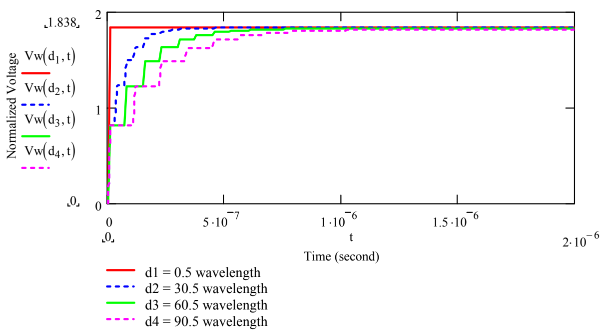

Figure 3 – Peak standing wave voltage vs. time in the transmission line for a reactively tuned case with a detuning $\Delta f = BW$ where $BW = fo/Q$

4 DISCUSSIONS

Using a transmission line T-junction (coaxial T or rectangular waveguide T) with one arm short circuited through a variable phase shifter in series can realize the variable reactance: H-plane T for a shunt stub and E-plane T for a series stub in rectangular waveguides.

For a cavity with a detuned frequency range of $\pm\Delta f$ where Δf is the half power bandwidth of a cavity, normalized shunt susceptance (series reactance) required for the tuning is $-1.4 < B(X) < 1.4$. Then, the equivalent short circuit transmission line length is $0.15\lambda < s < 0.037\lambda$ or the needed phase shifter phase range becomes $-54.2° < \phi < 54.2°$. This detuning compensation system could be viewed as a tuning system to change the resonance frequency of a tuned cavity.

The variable reactance for tuning a high power accelerating cavity may employ a fast phase shifter such as a ferrite phase shifter that can have fast response with high power handling capability. For fast response and lower loss operation, non-reciprocal phase shifters are often used. However, this may not be desirable, since the effective phase shifting may be decreased. For a non-reciprocal transmission line, the standing wave formed by the forward and reflected waves are

$$V(z) = V^+ e^{-\gamma_1 z}(1 + \Gamma_1 \; e^{(\gamma_1 + \gamma_2)z})$$

where Γ_l is the reflection coefficient at the load and γ_1 and γ_2 are the forward and reverse propagation constant, respectively. The effective wavelength on the standing wave will be $(\lambda_1+\lambda_2)/2$. If a non-reciprocal phase shifter has phase shifts ϕ_1 and ϕ_2 in forward and reverse directions, respectively, the effective range of phase shifting becomes $(\phi_1+\phi_2)/2$.

This frequency control appears to be sufficiently independent of the usual phase and gradient controls that it can be appended to existing RF controls with no special modifications.

Complementary to frequency control is the possibility of control of the external Q of the cavity. It is desirable to operate cavities at the highest controllable gradient, but Lorentz detuning then shifts the cavity frequency away from the desired RF frequency. Krejcik noted that it is desirable to control this parameter as well as the frequency, and feasible to do both with waveguide stubs. With Lorentz detuning sufficiently severe that the cavity falls outside the resonance band when the cavity is turned off, the RF drive frequency and power must be swept together to bring the cavity back to proper gradient (self-excited loop control, for instance), or a mechanical tuner must be actuated to sweep the cavity through resonance as the power is raised.

However, if one places a variable reactance on the input waveguide at the proper distance from the cavity, the external Q of the waveguide/cavity system can be altered to broaden the resonance while the power is raised. This opens the highly desirable option of active control of the external Q to optimize the coupling of the RF source to the cavity across a broad range of beam current. Significant reductions in power consumption are possible, especially using the frequency compensation to enhance microphonic control to allow operation of the RF system at external Q values formerly infeasible because of microphonic limitations.

5 ACKNOWLEDGEMENT

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6 REFERENCES

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