Non-Accelerator Applications of Rf Superconductivity

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I. INTRODUCTION

Particle accelerators for Elementary Particle or Nuclear Physics were from the beginning of the study of superconductors in high frequency fields the prime target of application and the main origin of funding for our field. It is, however, interesting to note that one of the first important applications of rf superconductivity was demonstrated by the initial operation of a Free Electron Laser at Stanford [1], an application which does not point in the direction of High Energy Physics. The application of a superconducting cavity in an oscillator of extremely low phase noise, the SCSO at Stanford gives another excellent example of a non accelerator application. The progress in superconducting FEL's is discussed in A. Schwettman's contribution to this workshop and the work on the Superconducting Cavity Stabilized Oscillator is published since long [2]. The last review on non-accelerator applications of superconducting cavities [8], does not yet contain the use of a superconducting cavity in a 'Single Atome Maser' [4] where the interaction of a single atom with one mode of a superconducting cavity at 21.5 GHz comprises an interesting experiment for the study of basic maser interactions. This experiment was already mentioned in the last workshop at CERN [3, 6]. A similar experiment at 68.4 GHz is presently under preparation and its superconducting cavity work is reported at this workshop [5]. The discovery of the high T_c superconductors has stimulated a new discussion about various potential applications. In our contribution we specifically want to concentrate on a special class of application to passive Rf devices and systems [7,8]. Electronic components (Josephson devices etc.) are not considered.

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Whereas the surface resistance R_s of normal conducting (n.c.) copper at room temperature is given by

$$R_s^{nc} = R_s^{nc}(f_o)\left(\frac{f}{f_o}\right)^{0.5} \quad \text{with } R_s^{nc}(f_o = 3\text{ GHz}) \approx 15\text{ m}\Omega \qquad (1)$$

(f: frequency, f_o: reference frequency)

$$R_s^{sc} \approx R_s^{sc}(f_o)\left(\frac{f}{f_o}\right)^2 \qquad (2)$$

holds for superconducting (s.c.) materials at a given temperature. Taking f_o = 3 GHz as reference frequency the surface resistance is reduced relatively to copper at room temperature for niobium at T = 4.2 K by a factor of about 10^{-4} and for $Y_1 Ba_2 Cu_3O_{7-\delta}$ at T ≈ 77 K by a factor of about 10^{-2} [9].

As shown in the next sections the very low value of the s.c. surface resistance may lead to significant improvements if at least one of the following properties is of prime importance:

1) Frequency response = frequency dependence of impedance and/or transfer function of a device.
2) Efficiency = ratio of 'used' power (with respect to the particular function of the device) to total rf input power.
3) Signal-to-noise ratio.
4) Sensitivity of measurement system for material parameters.

If a device has to meet certain requirements with respect to one or more of these properties, one could be faced with a situation where normal conducting (n.c.) structures are insufficient and one has therefore to take s.c. as a possible solution into account. There are two cases which should clearly be distinguished:

(a) Extreme properties (e.g. filters with extremely small relative bandwidth) which in no case can be achieved with normal conducting devices.

(b) 'Normal' properties which can be achieved in the decimeter-wave region with normal conducting devices of relatively 'large' size but which are required to be offered by a miniaturized components in this frequency region or by millimeter- and submillimeter-devices.

II. COMPONENTS WITH SPECIAL REQUIREMENTS REGARDING THEIR FREQUENCY RESPONSE

General consideration:

A wide class of rf-components such as nondispersive and dispersive transmission-lines, filters etc. are designed to produce the desired transfer function as a non-dissipativ reciprocal two-port between resistive terminations. Their properties may be represented by the complex valued transmission coefficient $S_{21}(f)$, which relates the outcoming wave at port 2 to the incident wave at port 1 (magnitude and phase, time dependence exp $(j\omega t)$). With

$$S_{21}(f) = \exp\left[-a(f)-j\varphi(f)\right] \tag{3a}$$

a(f) denotes the insertion loss, while the group delay T_{gr} is given by

$$T_{gr}(f) = \frac{1}{2\pi}\frac{d\varphi}{df}\,. \tag{3b}$$

Since in the ideal non-dissipativ case no power is absorbed within the two-port, $a(f)>0$ is due to the fact that power is reflected back to the source. But because all real components exhibit unwanted dissipation due to metallic, dielectric and sometimes radiation losses a(f) and $T_{gr}(f)$ differ from the (theoretical) functions for lossless structures. In many components this difference does not considerably affect the function, but in some components dissipation can lead to a severe and therefore not tolerable degradation in system performance. To obtain a crude first approximation about the order of degradation by loss effects the 'equal uniform dissipation approximation [10]' can be utilized.

With this approximation, where the unloaded quality factor Q of all elements composing the two-port is taken to be equal, the insertion loss $\tilde{a}(f)$ of the lossy two-port turns out to be related to the insertion loss a(f) and group delay $T_{gr}(f)$ of the lossless two-port by

$$\tilde{a}(f) = a(f) + 2\pi f\, T_{gr}(f)\,\frac{1}{Q}\,. \tag{4}$$

This leads to the important conclusion that the influence of dissipation losses onto the insertion loss is 'amplified' by the group delay T_{gr}. Hence, components with a large value of $f\cdot T_{gr}$ are extremely 'sensitive' to dissipation. As examples for such components narrow bandwith filters, pulse compression filters, dispersion-free transmission lines for very short impulses and cavities for oscillator-stabilization are considered.

The (unwanted) dissipation losses in a component of the two-port are quantitatively represented by the unloaded quality factor

$$Q_o(f) = 2\pi f \frac{W}{P} , \tag{5}$$

where W denotes the maximum stored energy. The average dissipation power P is in general the sum of (metallic) wall losses P_W and dielectric losses P_d . In the case of open structures (e.g. microstrip circuits) additional radiation losses P_{rad} occur. With

$$\frac{1}{Q_o} = \frac{1}{Q_W} + \frac{1}{Q_d} + \frac{1}{Q_{rad}} \tag{6}$$

where Q_W, Q_d and Q_{rad} belongs to wall, dielectric and radiation losses respectively, a necessary condition for a useful application of superconducting ($Q_W = Q_W^{sc}$) instead of normal conducting ($Q_W = Q_W^{nc}$) wall material is the dominance of the n.c. wall losses:

$$Q_W^{nc} \ll Q_d \quad \text{and} \quad Q_W^{nc} \ll Q_{rad} . \tag{7}$$

Q_W of a component (n.c. or s.c.) is proportional to Z_o/R_s and is monotonically increasing with the relative size D/λ (D: typical linear dimension, $\lambda = c/f$, Z_o = 377 Ω):

$$Q_W = \frac{Z_o}{R_s} F\left(\frac{D}{\lambda}\right) \quad \text{with} \quad \frac{d}{dx} F(x) > 0 . \tag{8}$$

The meaning of the parameter D/λ is illustrated in Fig. 1 , where a resonant circuit serves as example. For $D/\lambda \approx 0.5$ it is realized by means of a fundamental mode cavity and for $D/\lambda >> 0.5$ by a (oversized) higher order mode cavity. To reduce the relative size below 0.5 (miniaturization) a stepped inpedance coaxial resonator ($D/\lambda \approx 1/8$) or a lumped element circuit may be utilized.

For the purpose of a principal discussion of the size dependence the function $F(D/\lambda)$ in eq. (8) may be approximated by

$$F(D/\lambda) \approx F_o \cdot (D/\lambda)^{\nu} \quad \text{with } 1 \le \nu < 3, \tag{9}$$

where the constants F_o and ν depend on the particular structure of the circuit. If for a component (e.g. the resonant circuit of a filter) Q_W is specified according to the considerations given above, for n.c. circuits

$$\frac{D}{\lambda} > \left(\frac{Q_W}{F_o} \frac{R_s^{nc}(f_o)}{Z_o}\right)^{\frac{1}{\nu}} \left(\frac{f}{f_o}\right)^{\frac{1}{2\nu}}$$

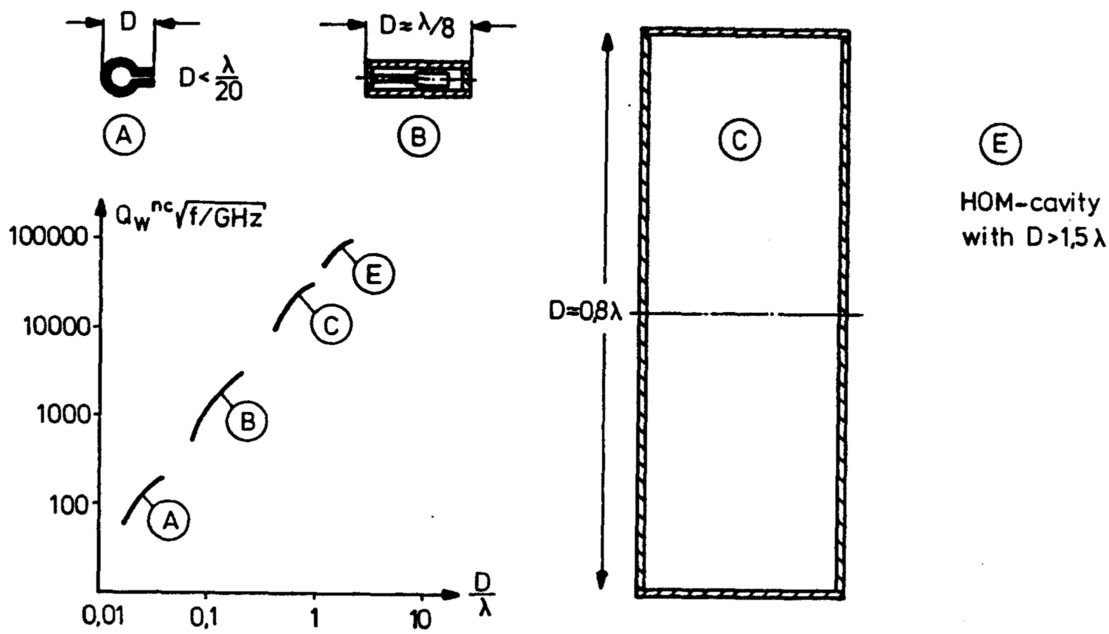

Fig. 1:
Realization of a resonant circuit for the same resonance frequency by means of components with different relative size. A: lumped element circuit, B: stepped impedance coaxial resonator, C: fundamental mode cavity, E: Higher order mode cavity. Curve at the left part shows the corresponding unloaded Q factors for the normal conducting case.

and for s.c. circuits

$$\frac{D}{\lambda} > \left(\frac{Q_W}{F_o}\,\frac{R_s^{sc}(f_o)}{Z_o}\right)^{\frac{1}{\nu}} \left(\frac{f}{f_o}\right)^{\frac{2}{\nu}}$$

is required. Fig. 2 indicates the principal meaning of this requirement using $Q_W/F_o = 10^4$, $\nu = 1$ and R_s^{sc} (3 GHz) $\approx 10^{-2}$ R_s^{nc} (3 GHz) as numerical example. For f – (D/λ_o) – combinations belonging to region I this requirement can be met with normal conducting material. In region II superconducting material must be used and in region III the requirement can not be fulfilled.

From Fig. 2 three distinct cases where superconductivity has practical importance can be deduced:

(A) Structures (relative size $D/\lambda \approx 0.5$, non-overmoded) in the millimeter – and submillimeter – wave region with moderate Q_W values ($Q_W \approx$ 1 000 – 10 000) (see point A in Fig. 2).

(B) Miniaturized structures ($D/\lambda << 1$) for 'relatively low' frequencies (e.g. decimeter waves) with normal Q values (point B in Fig. 2) have to be taken into account, if a size in the order of $\lambda/2$ is not permitted.

(C) Structures with extremely high Q values ($Q > 10^5$) which in no case can be achieved with normal conducting devices. It is necessary to realize these structures in decimeter wave region with $D/\lambda \geq 0.5$.

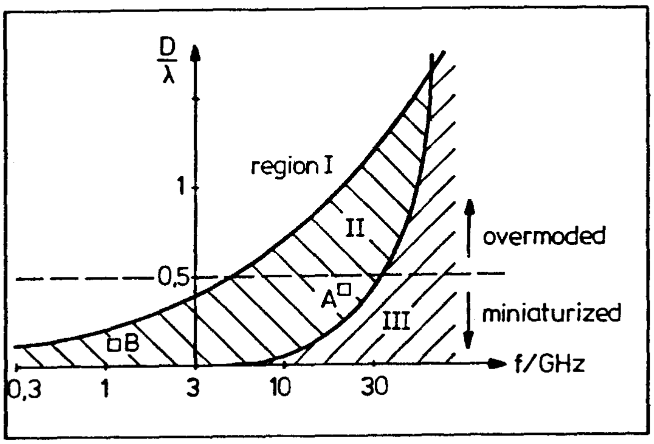

Fig. 2: Realizibility of a device with a given Q_W value as a function of frequency and relative size D/λ. Requirements can be fulfilled with normal conducting material in region I, with superconducting material in region II and can not be fulfilled in region III.

Bandpass filters with narrow band width:

From system theoretical considerations one finds that in the neighbourhood of frequencies where $|da/df|$ obtains large values (transition region) $T_{gr}(f)$ becomes very high, too. Therefore dissipation leads to strong degradations near the transition region.

For a bandpass filter composed by N resonators the insertion loss at center frequency f_o which would be zero without dissipation becomes (see also Fig.3)

$$\frac{\tilde{a}_o}{dB} \approx 4.34 \frac{f_o}{\Delta f} \sum_{n=1}^{N} \frac{g_n}{Q_{on}} . \qquad (10)$$

Here the Q_{on} are the unloaded quality factors of the resonators and the coefficients g_n depend on the filter type ('shape' of the frequency response) but are independent of Δf [11,12]. If both a low value of $\tilde{a}_o$ and an extremely small bandwith $\Delta f/f_o$ are required Q_{on} values which cannot be obtained with normal conducting structures may result (example: $\tilde{a}_o \leq 1$ dB, $\Delta f/f_o = 2 \cdot 10^{-5}$ results in $Q_{on} \approx 430\ 000$).

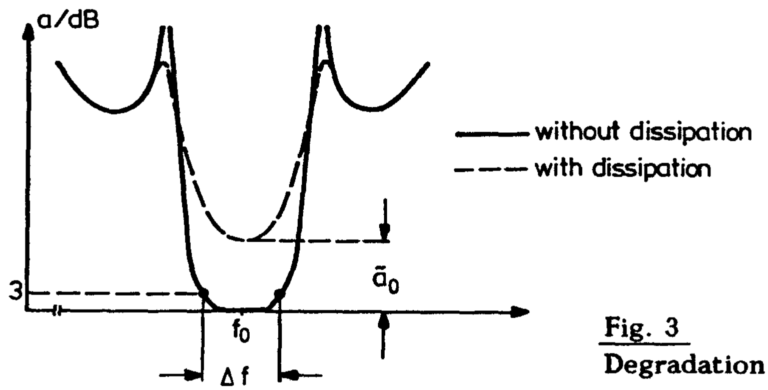

Fig. 3
Degradation of the frequency response of a bandpass filter due to dissipation loss. ($\tilde{a}_o$: insertion loss at center frequency f_o).

Possible applications of filters with extremely low values of the product $\tilde{a}_o$; $\Delta f/f_o$ are (1) receiver preselector filters to avoid intermodulations in the rf stages by a strong signal which is closely spaced in frequency [13].
(2) phase noise reduction in multiplier type microwave sources.
These oscillators use multiplier chains to generate stable microwave signals from quartz crystal stabilized sources (with e.g. f = 5 MHz). One disadvantage of this principle is that phase noise is increased by multiplication. The SSB phase noise to carrier ratio is increased by 20 log n after a frequency multiplication by n [14]. To reduce this effect an extremely narrow bandpass filter could be used to narrow the spectral width (and therefore the phase noise) in the microwave region [15].
Some results on superconducting filters may be found in [13, 16-18]. The experimental results reported in [17] turned out to be wrong due to measurement errors [18].

Pulse compression filters:

In radar techniques frequency modulated (FM) pulses ('chirp') with a large time bandwith product $\tau \cdot \Delta f$ are used because they provide both good range resolution and high energy [19]. At the receiver a pulse compression (p.c.) filter with a dispersive group delay characteristic that has an inverse relationship to the FM-function is needed. P.c. filters can be realized in a number of different ways, e.g. with ultrasonic delay lines, optical signal processors etc. [19]. If an extremely wide bandwidth is required a microwave p.c. filter (folded type meander line as slow wave structure [19,20], tapped delay line [21]) may be a proper design.

For large compression ratios the proper differential group delay $T_{gr,max}-T_{gr,min}$ is in the order of the time duration ΔT of the uncompressed pulse, so that for the maximum group delay $T_{gr,max} > \Delta T$ must hold.

From eq. (4) the maximum insertion loss due to dissipation - being largest where the delay T_{gr} is maximum - turns out to be

$$\tilde{a}_{max}/\mathrm{dB} \approx 5.5 \cdot 10^4 \cdot \frac{f_o}{\mathrm{GHz}} \cdot \frac{\Delta T}{\mu s} \cdot \frac{1}{Q} \qquad (11)$$

A typical p.c. filter with a center frequency of f_o = 5 GHz for frequency modulated pulses with a duration $\Delta T = 1$ µs and a bandwidth Δf = 1 GHz may have a compression ratio of 1000 leading to a compressed duration $\Delta T' = \Delta T/1000 = 1$ ns. For this example a Q value greater than 27 000 is required from eq. (11) if $\tilde{a}_{max}$ is not allowed to exceed 10 dB. Because this low dissipation cannot be obtained with normal conducting structures pulse compression filters are potential applications of superconductivity. Gondolfo et al. [22] demonstrated this possibility with a superconducting meander line [lead, T = 4.2 K] and achieved less than 16 dB/µs insertion loss.

Dispersion free transmission line for very short impulses:

In some applications very short impulses (fast transients) have to be transmitted over relatively large distances. If the transfer function of the transmission line is except for a linear phase term frequency dependent (dispersive) the temporal waveform is distorted (e.g. rise time altered to a lower value) along the line. This leads in certain applications to a degradation of system performance. In a digital circuit e.g. it results in speed limitations. An ideal distortion free transmission line requires the phase velocity and the attenuation of the line to be independent of frequency. If (in theory) lossfree transmission lines are considered one finds that only TEM modes have a constant phase velocity. But if for the TEM mode dissipation loss is also taken into account a frequency dependent attenuation and group delay results which leads to pulse distortions.

If the input signal to a lossy n.c. TEM line is a step function the output signal will due to distortion exhibit a nonzero rise time T_r given by

$$T_r \approx 128\ \tau \left(\frac{l}{D}\right)^2 \Lambda^2 \qquad (12)$$

with $\tau = \varepsilon_o/\varkappa \approx 1.7 \cdot 10^{-7}$ ps for copper. D denotes a typical transverse dimension and l the length of the line whereas Λ is a shape factor. For a coaxial line with outer and inner diameter of the conductors given by D and $d = D/3.6$ respectively Λ = 1 results.

In order to avoid higher order mode propagation for the transverse dimension

$$2\ D/\lambda_{min} = \eta < 1 \quad (\eta : \text{miniaturization factor}) \qquad (13)$$

must hold. If the transmission of an impulse with a time duration ΔT is considered and the crude approximations $T_r < \Delta T$ and $f_{max} = c/\lambda_{min} = \pi/\Delta T$ are introduced into eqs. (12) and (13) one obtains for a normal conducting copper transmission line at room temperature

$$\frac{l_{max}}{m} \approx \frac{\eta}{100\Lambda} \left(\frac{\Delta T}{ps}\right)^{1.5} \qquad (14)$$

l_{max} is the maximum allowed length with respect to impulse distortion. If for very short impulses (e.g. rise time in the order of 10 to 100 ps) the maximum length ($l_{max} \approx 0.3$ to 10 m) obtained from eq. (14) for n.c. lines is not sufficient for the particular application, a superconducting line may be taken into account. Nahman [23] investigated the possibility of a superconductive coaxial transmission line (1.6 mm diameter, T = 4.2 K, Pb conductors) as a delay line for nanosecond pulses. At 1 GHz the losses predominantly caused by dielectric losses leading to an attenuation in the order of 1 dB/km. Mikoshiba et al. [24] also deal with a s.c. coaxial line. They point out that a difficult problem arises from periodic impedance irregularities resulting due to multiple reflections in high nondissipative attenuation for particular frequencies. McCaa and Nahmann [25] as well as Voges and Petermann [26] have calculated losses in superconducting transmission lines. Since it can be expected that the difficulties due to dielectric losses and multiple reflections can be overcome, transmission lines for very short impulses seem to be important applications of superconductivity.

Superconducting cavity-stabilized oscillators:

All oscillators exhibit unwanted random phase fluctuations (phase noise). This short term frequency instability is commonly characterized by a function which describes the spectral density of phase noise as a function of the offset from carrier frequency (SSB (single side band) phase to noise carrier ratio) [27]. Some applications have such strong phase noise requirements that not even a synthesized microwave source can satisfy its requirements.
Examples:
In a Doppler radar typically a 'small' echo from the desired moving target which is shifted in frequency according to its velocity has to be separated from strong echos of undesired stationary objects (clutter). In this situation the ability of a radar to resolve targets with a small Doppler shift is limited by the phase noise of either the transmitter or the receiver local oscillator.
In digital communication systems phase modulation is often used to transmit digital signals. Here the phase shift of the carrier is switched between different values which are related to different 'numbers' to be transmitted. As a result of local oscillator phase noise the digital phase detection makes an incorrect decision if the phase error exceeds the minimum phase step of the modulation.

A cavity with a high Q factor may be used in different ways to stabilize oscillators. It can be employed to built a frequency discriminator for an automatic phase control loop (cavity, balanced mixer, phase modulator). In this case the phase noise reduction is mainly determined by the noise level of the control loop and the unloaded Q-factor of the resonator.

The resonator may also be used as a 'transmission stabilizing cavity [28]', where the cavity serves both as frequency dependent susceptance stabilizing the oscillator proportional to the unloaded Q-factor and as a very narrow-band bandpass filter leading to a further noise reduction.

A number of papers dealing with superconducting cavity stabilized oscillators (see also [8]) exist. Stein and Turneaure in 1972 [29, 30] reported very good short-term and medium-term stability ($\sigma_y \sim 10^{-15}$ in the time interval 10-1000s).

III. EFFICIENCY ENHANCEMENT BY S. C.

Accelerator cavities:

The ratio of power delivered to the beam ('used power') to the total rf power fed into the cavity is dramatically increased by using s.c. Since this application is discussed elsewhere it is not further considered here.

Electrically small and superdirective antennas:

An antenna may be considered as a 'mode transformer' which provides the transition from the guided wave in the transmission line to the radiating modes in free space. If the geometrical size of an antenna is small compared with the free space wavelength λ ('electrically small') the 'coupling' between the waveguide mode and the radiating modes is 'very small', too. A parameter which describes this coupling quantitatively is the radiation resistance R_{rad} defined by

$$P_{rad} = 0.5\ I^2\ R_{rad}, \qquad (15)$$

where I denotes the magnitude of the rf current flowing into the antenna and P_{rad} denotes the radiated power. For a dipole with length l much less than λ

$$R_{rad} \sim (l/\lambda)^2 \qquad (16a)$$

and for a small loop antenna with loop area A

$$R_{rad} \sim (A/\lambda^2)^2 \qquad (16b)$$

holds.

Since furthermore a large amount of stored energy W is associated with an electrically small antenna the equivalent circuit (see Fig. 4) is given by a low

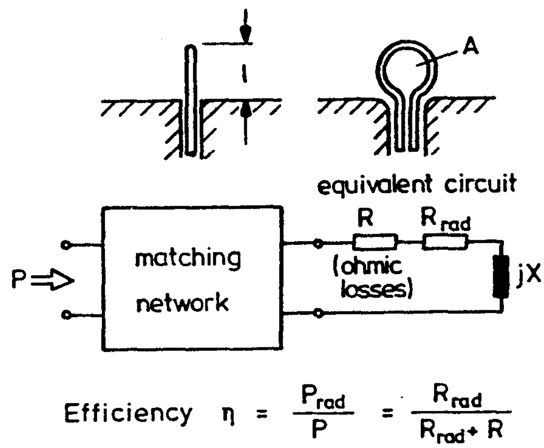

Fig. 4:
Electrically small antennas (dipole and loop antenna) with corresponding equivalent circuit.

radiation resistance in series with a high reactance (capacitive for the dipol and inductive for the loop antenna). The ohmic resistance of the antenna is represented by R. To provide maximum power transfer into the antenna a matching network has to be inserted between the transmission line and the antenna. At room temperature the ohmig resistance is much larger than the radiation resistance. That means that the power fed into the antenna is mainly dissipated and only a 'small' amount is radiated. So the efficiency can approach very low values. With a superconducting antenna R can be reduced to be less than R_{rad}. Walker and Haden [31] studied experimentally a s.c. loop antenna. Whereas the efficiency at room temperature was about 1% they obtained an efficiency of nearly 100% by cooling down the structure (antenna and matching network) to 4.2 K.

Besides the input impedance another important property of antennas is their ability to concentrate radiated energy in particular directions (directional antennas). This property is quantitatively described by the directivity which is the ratio of the radiation intensity in the main direction and the mean radiation intensity. An array is a special kind of a directional antenna. It is a group of single antennas (radiation elements), which are coherently fed from a single source. Assuming for the sake of simplicity that the single elements radiate isotropically the radiation pattern is produced due to the interference of the contributions of all elements. The angle dependency of the radiation pattern is based on the fact that the differences in the path lengths from the elements to a point in the far field change with the angle. Angles where the contributions add constructively or destructively can be influenced by the amplitudes and phases of the currents which are fed into the radiating elements. These amplitudes and phases are determined by the associated branching network (see Fig.5). If the elements are situated along a line with fixed spacing (total

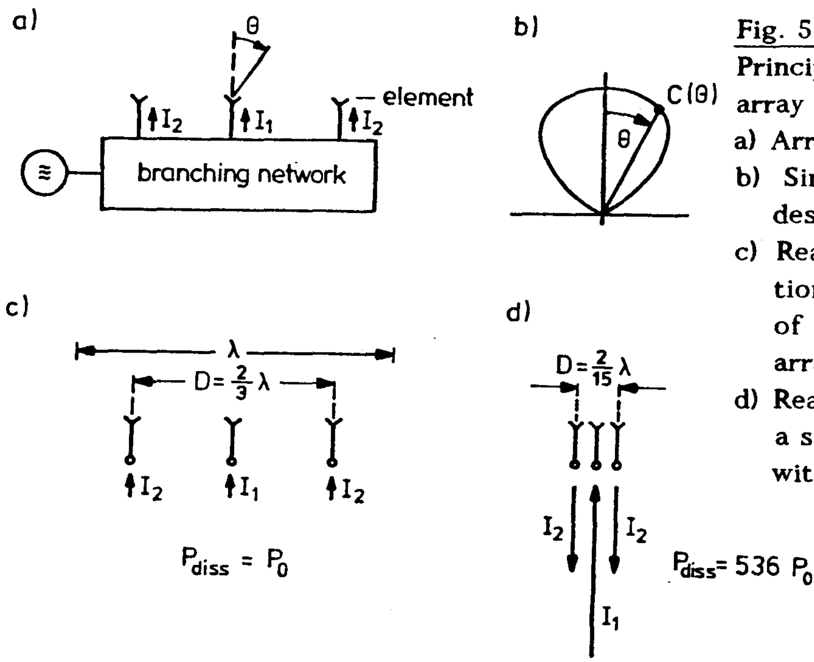

Fig. 5:
Principle of a superdirective array
a) Array with 3 elements
b) Simple example for a desired radiation pattern
c) Realization of the radiation pattern by means of a normal directive array with $D=2\lambda/3$
d) Realization by means of a superdirective array with $D=2\lambda/15$.

length of the array $D \gg \lambda$) and all elements are fed with currents equal in phase and magnitude a high directive pattern with a 3 dB beam width $\Delta\Theta$ given by

$$\sin(\Delta\Theta) \approx \lambda/D$$

and a main beam directed orthogonal to the array results. From this equation one could conclude that a high directivity can only be realized with an antenna being large in comparison to the wavelength. But this conclusion is not true because one can (in theory) obtain any desired directivity value with an antenna of any size and therefore also with an array of transverse dimension $D \ll \lambda$ [32]. An array with a directivity superior to the directivity of an array with equal currents in all elements is called a superdirective array [32,33].

The principle of a superdirective array may easily be explained by means of the array shown in Fig. 5a. It is composed of 3 isotropic radiating elements. For sake of simplicity it is assumed that a radiation pattern like the pattern shown in Fig. 5b is to be realized.

From a simple interference consideration the radiation pattern $C(\Theta)$ is found to be given by

$$C(\Theta) = \text{const} \cdot |I_1 + 2\, I_2 \cos(\pi D \sin\Theta/\lambda)|^2 .$$

To obtain $C(\Theta = \pm\pi/2) = 0$ the currents have to be chosen according to

$$I_2 = -I_1 / (2\cos(\pi D/\lambda)).$$

The last equation shows that one can find a solution for any D.

(α) For $D = 2\lambda/3$ one obtains $I_2 = I_1$, leading to a normal directive antenna(see Fig. 5c). The radiation intensity for the direction $\Theta = 0$ is given by

$$C(0) = \text{const.} \cdot 9\, |I_1|^2.$$

(β) For $D = 2\lambda/15$ one obtains $I_2 = -0.547\, I_1$ (Fig. 5d) and

$$C(0) = \text{const.} \cdot 0.009\, |I_1|^2.$$

Hence the current $|I_1|$ has to be chosen 31.7 times greater for case (β) as in case (α) to obtain the same radiation intensity.

Figs. 5c and 5d illustrate this result and indicate the typical feature of super-directive antennas. They utilize large and alternating element currents. If the ohmic resistance is assumed to be the same in each of the 3 elements the ohmic losses are in case (β) about 536 times as high as in case (α). That clearly shows the advantage which could be offered by a superconducting structure [34].

If the dimension D is kept constant, but more than 3 elements are used the directivity and therefore also the beam width can (in theory) be arbitrarely increased. But in practice close constraints have to be taken into account due to the
(1) sensitivity of the array to small errors in the magnitude of the currents and geometrical parameters (e.g. distance between elements).
(2) reduced frequency bandwith.

Electrically small antennas and superdirective antennas are of interest to those applications where antenna dimensions comparable to the wavelength are unaccep-table.

IV. IMPROVEMENT OF SIGNAL-TO-NOISE RATIO WITH S.C. PASSIVE COMPONENTS

The weakest signal that can be detected by a receiving system is usually determined by the amount of noise which is accompanying the signal. This noise has its origins in the noise present at the input of the receiving system as well as in the noise which is generated within the system itself. The latter contribution is described by the noise temperature T_{eff} (in K) or the noise figure

$$F = 1 + T_{eff} / 290 \quad .$$

If the receiving system is composed of a passive network (transmission lines, power dividers, switches etc.) and a low-noise amplifier the overall noise properties may be considerably degraded by the losses in the passive network. To study this effect and the improvements obtainable by cooling the passive network a cascade of this transmission line and an amplifier is considered (Fig. 6).

If the effective noise temperature (in K) of the amplifier is denoted by T_a, the (physical) temperature (in K) of the transmission line by T and the loss factor by L, with

$$L = \exp(2\alpha l)-1 \qquad (l = \text{line length, } \alpha = \text{attenuation constant})$$

one obtains

$$T_{eff} = LT + (L+1)T_a \quad . \tag{17}$$

The degradation of the signal-to-noise ratio (SNR) is given by

$$(SNR)_2 = \frac{1}{1 + T_{eff}/T_{in}} (SNR)_1 \quad ,$$

where T_{in} is the noise temperature of the source connected to the input of the receiving system (e.g. antenna temperature).
The first term in eq.(17) is due to the contribution by the lossy line itself whereas the second term represents the contribution by the amplifier increased by the attenuation in the transmission line. Equation (17) indicates that a reduction of the noise temperature can be attained by reducing the (dissipative) loss factor L or the physical temperature T of the transmission line or by a combination of both measures.

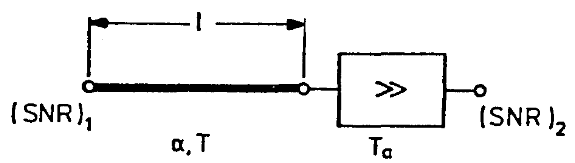

Fig. 6:
Configuration consisting of a transmission line and an amplifier.
$(SNR)_1$, $(SNR)_2$: input and output signal-to-noise ratios

Cooled front end of a high sensitive receiver:

In certain applications as e.g. in radio astronomy the received signal level is extremely low. Therefore cryogenically cooled low noise amplifiers (e.g. Maser) are used. To avoid a degradation of the SNR due to large transmission line networks (especially with array antennas) and other passive components (switch etc.) between the antenna and the first amplifier ('front end') these part of the system is cooled, too. Because this part is usually cooled to a temperature T of about 20 K the components are still normal conducting. Therefore the quantity L in eq. (17) which represents absorption loss is not extremely reduced and the improvement in noise figure is due to the fact that T is an extra factor in eq. (17).

With a new hight T_c-superconductor a much larger noise reduction could perhaps be obtained with a temperature of about $T \approx 80$ K, because in this case both L and T decrease considerably.

Long distance s.c. communication cable:

In a communication system the signal-to-noise ratio at the end of a transmission medium e.g. a cable, has to exceed a certain level to assure that the information contained in the signal can be extracted. A given value of $(SNR)_2$ determines the maximum length of a cable between two amplifiers. If one considers a s.c. cable as a long distance transmission medium, one has to take into account that such a cable has to compete with an optical fiber line. Because with optical fibers attenuation as low as 0.15 dB/km are attained there seems to be no need for this application of superconductivity.

V. SENSITIVITY OF MEASUREMENT SYSTEM TO MATERIAL PARAMETERS

A well known measurement technique is to insert a small sample of a material into a cavity and to determine the dielectric (complex permittivity) and/or magnetic (complex permeability) of the material from the shift in resonance frequency and quality factor of the cavity. The use of s.c. resonators instead of n.c. resonators at room temperature increases the sensitivity of the method by several order of magnitude [35-38]. This enables to determine much smaller loss tangents (up to $\tan \delta = 10^{-9}$) as well as much smaller changes in the real part of the permittivity or permeability.

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