Electrical Mode Stirring in Reverberating Chambers by Reactively Loaded Antennas

Edgar Voges and Tobias Eisenburger

Abstract—Electrical mode stirring in reverberating chambers for electromagnetic compatibility is accomplished by antennas that are loaded by variable reactances. The principle is derived from a network model of the chamber that employs measurable coupling factors. The concept is experimentally confirmed in the gigahertz-range for a 1.92 m × 1.84 m × 2.91 m chamber with four conical antennas placed on one sidewall. One antenna is connected to a generator; the other three antennas are reactively loaded by variable short circuit plungers. The resonances are considerably shifted by varying the reactive loads.

Index Terms—Electrical mode stirring, electromagnetic compatibility (EMC), reactively loaded antennas, reverberating chambers.

I. INTRODUCTION

MODE-STIRRED or mode-tuned reverberating chambers are now an important tool for electromagnetic compatibility (EMC) measurements. The modes of a large overmoded cavity are perturbed in resonant frequencies and field distributions by a relatively large metallic stirrer to achieve, on average, an isotropic and homogeneous electromagnetic field. The basic electromagnetic and statistical properties of such mode-stirred/tuned reverberating chambers are well documented [1]–[5].

Mode-stirred/tuned reverberating chambers for EMC are standardized by several organizations, e.g., by the standard IEC 61000-4-21:2003.

Therefore, several alternative techniques have been proposed as a replacement for the mechanical stirrer. Electronic mode stirring has been evaluated in [6]; it may be used in addition to other techniques. A moving antenna (source stirring) has been investigated in [7]. An excitation by antenna arrays has been investigated in [6] and [8] with simplified models.

In this paper, we investigate the novel mode stirring technique by antennas, which are loaded by variable reactances. Variations of the reactive loads change the resonance frequencies and provide mode stirring. This technique is described in Section II by a network model of the reverberating chamber with antennas. The function of the antennas is given in terms of coupling factors of the antennas to the resonances, which are related to external quality factors. This model points out the method of operation and allows basic design considerations.

The experimental investigations in Section III are performed for a 1.92 m × 1.84 m × 2.91 m chamber with copper walls and unloaded quality factors above 10⁵. Four conical antennas, designed for frequencies beyond 500 MHz, are placed on one sidewall. One antenna is the transmitting antenna; the three other antennas are reactively loaded by variable coaxial short-circuit plungers. Precise measurements of the changes in resonance frequencies and of the coupling factors are obtained with a calibrated network analyzer.

The results show considerable shifts of the resonances when changing the variable short-circuit plungers. The frequency shifts are random for random variations of the plungers.

Furthermore, a basic statistical description is presented by calculating the scatter plot and the autocorrelation function of the transmission coefficient of the transmitting antenna and a receiving antenna. Detailed investigations of the achievable field homogeneity will follow.

II. PRINCIPLE AND NETWORK MODEL

Fig. 1(a) shows the basic configuration of an overmoded cavity (chamber) with two antennas. One antenna is connected to a generator, and the second antenna is terminated by a variable reactance X. We define reference planes in the connecting coaxial lines of wave impedance Z₀ = 50 Ω.

The basic approach is to treat the antennas as coupling elements to the cavity. Then, we can adopt network models and calculation techniques from the theory of cavities coupled by, for example, probes, loops, or apertures [9]–[11]. A simplified network model for the arrangement of Fig. 1(a) is depicted in Fig. 1(b) [9], [10]. It consists of infinitely many parallel resonant circuits with resonant frequencies f_k and resonance resistances R_k connected in series. We assume that all resonant circuits have the same value

\[ \sqrt{L/C} = Z₀ = \sqrt{\mu₀/\varepsilon₀} = 120/π \Omega. \]  \hspace{1cm} (1)

(A description with admittance parameters and a network with series resonant circuits connected in parallel would basically be equivalent. The choice of the model depends on the type of coupling. A quarter-wavelength shift of the input reference plane would—anyway—change a parallel resonance to series resonance and vice versa.) In this network model, we omit connecting transmission lines [Fig. 1(a)] and the antenna reactances. One recognizes the possibility of electrical mode stirring by reactively loaded antennas. The variable reactance jX is transformed with jX/\(n_{jk}^2\) into the resonant circuits and changes their resonant frequencies f_k.

The transformer ratios \(n_{jk}\): 1 with the first index \(i = 1, 2\) in our case indicating the antenna and the second index \(k = 1, 2, 3, \ldots\) indicating the resonances can be expressed by measurable coupling coefficients \(K_{jk}\). The \(K_{jk}\)'s are related to

Manuscript received January 19, 2007; revised July 4, 2007. This work was supported by the German Research Council (DFG) under Project Vo 200/38.

The authors are with the High Frequency Institute, University of Dortmund, D-44221 Dortmund, Germany (e-mail: edgar.voges@udo.edu; tobias.eisenburger@udo.edu).

Digital Object Identifier 10.1109/TEMC.2007.908281
The two-port 1, 2 is described by impedance parameters according to
\[
U_1 = z_{11}I_1 + z_{12}I_2 \\
U_2 = z_{21}I_1 + z_{22}I_2
\]
with \(z_{12} = z_{21}\) for a reciprocal two-port. A generalization to more than two coupling antennas, for example, to \(n\)-ports, is easily obtained.

The impedance parameters \(z_{11}\) and \(z_{22}\) are given by
\[
z_{11} = \sum_k \frac{1}{Y_{1k}} \\
z_{22} = \sum_k \frac{1}{Y_{2k}}.
\]

The impedance parameter \(z_{21}\) represents the coupling between both antennas. Here, we have to introduce coupling parameters
\[
K_{i'i''k} = \sqrt{K_{ik} \cdot K_{i''k}}, \quad \text{with } i', 2, \text{ and } i'' \neq i
\]
and obtain the coupling admittances
\[
Y_{i'i''k} = \frac{1}{K_{i'i''k}Z_L} \left\{ j \left( \frac{f}{f_k} - \frac{f_k}{f} \right) Q_{0,k} + 1 \right\}
\]
\[
z_{21} = \sum_k \frac{1}{Y_{i'i''k}}
\]

We observe from (3) and (7) that the impedance parameters reach appreciable values only near the resonances.

With the aid of the impedance parameters the input reflection is given by \(r_{in}\) at port 1 with the input resistance \(Z_{in}\)
\[
Z_{in} = z_{11} - \frac{z_{12}^2}{z_{22} + jX} \\
r_{in} = \frac{Z_{in} - Z_L}{Z_{in} + Z_L}
\]

Fig. 2 presents one example for electrical mode stirring by a varying reactance \(jX\). We consider three resonances near 2000 MHz, \(f_k = 1999.83/1999.9/2000.13\) MHz with \(Q_{0,k} = 50000\), and coupling factors \(K_{11} = 0.4, K_{12} = 0.8, K_{13} = 0.6, K_{21} = 0.2, K_{22} = 0.4, \) and \(K_{23} = 0.8\). Three values of \(X\) are chosen: \(X \to \infty (Z_{in} = z_{11}), X = +5 \Omega, \) and \(X = -5 \Omega. \) One observes a considerable shift of the resonances.

With regard to measurements, the two-port is more adequately described by scattering parameters \((s\text{-parameters})\). Port 2 is then loaded by \(Z_L\). The impedance parameters are converted into \(s\text{-parameters} yielding for the transmission coefficient
\[
s_{21} = s_{12} = \frac{-2z_{21}Z_L}{Z_L + z_{11} + z_{22} + z_{11}z_{22} - z_{21}^2}.
\]

Fig. 3 shows \(|s_{21}|\) for the parameters given in Fig. 2. The \(s\text{-parameter} \ |s_{21}|\ reaches a maximum value of \(-3\) dB, if critical coupling of both antennas is established.

Generally, these simple calculations show that an efficient resonance shifting is obtained under the following conditions.

1) The coupling factors should be near critical coupling. Strong undercoupling or strong overcoupling should be avoided.
Fig. 2. Input reflection $|r_{in}|$ for three values of the reactance $X$ at port 2: $X \to \infty (Z_{in} = z_{11}), X = +5 \Omega, X = -5 \Omega$.

Fig. 3. Magnitude $|s_{21}|$ of the transmission coefficient for parameters of Fig. 2.

2) The reactances $|X|$ should be low, i.e., $|X| < Z_L$, e.g., one should operate near the short-circuit condition.

3) A short-circuit should, however, be avoided.

4) Parasitic resistances of the reactive loads must be low. The losses of the reactances should only be a few 0.1 dB.

III. EXPERIMENTAL RESULTS

The experimental investigations are performed with a reverberating chamber as shown in Fig. 4. The chamber with $1.92 \text{ m} \times 1.84 \text{ m} \times 2.91 \text{ m}$ volume is fabricated with soldered Cu-plates (1 mm thickness) employing outside mechanical stabilizing elements.

The door to the chamber is shielded by metallic mesh-surrounded elastomer tubes (Laird Technologies). The unloaded quality factor $Q_0$ is above 100 000 at frequencies near 1 GHz. The measured $Q_0$-values are not far below the theoretical values. (One should note that the quality factors in [1] do not include mode degeneracy. More accurate calculations according to [10] show considerable differences, which originate from a coupling of degenerated modes by conductive wall losses.)

Four conical antennas [13] are placed on one sidewall at arbitrarily chosen positions. The antennas are designed for 50-Ω impedance of the TEM transmission line mode, which requires according to

$$Z = 60 \Omega \ln \left\{ \cot \left( \frac{\theta_o}{2} \right) \right\}$$

a half cone angle $\theta_o = 46^\circ$ [9].

Fig. 5 shows a cross section of the antenna.

The height $h = 0.152 \text{ m}$ of the antennas made of 0.5-mm Cu-plates is chosen for operating frequencies above 500 MHz. All measurements are performed with the RF-network analyzer Agilent E5071B, which is calibrated for two-port measurements by the standard through/open/short/match (TOSM) method. Fig. 6
Fig. 6. Calculated and measured input reflection $|r_{in}|$ of the conical antennas for free-space radiation in the frequency range 0.1–8.5 GHz.

![Locus of the input resistance for a resonance at $f_k = 515.99$ MHz in the Smith chart. The ratio $R_{in}^k / Z_L$ at resonance determines the coupling factor; the unloaded quality factor $Q_{0,k}$ is determined from the bandwidth $2\Delta\omega_{0,k}$ between the locus and the circle segments $|X_k|^n = R_{in}^k$.](image)

shows the input reflection $|r_{in}|$ of the antennas for free-space radiation. The calculated values are obtained from the time-domain results employing the software package Microwave Studio [Computer Simulation Technology (CST)]. The measured values are obtained in a semianechoic hall. The input reflection is sufficiently low in the frequency range 1–4 GHz.

The unloaded quality factors $Q_{0,k}$ and the coupling coefficients $K_{1,k}$ or $K_{2,k}$ are determined by the standard technique of tracing the locus of the input reflections in a Smith chart, and determining $R_{in}^k$ at resonance and the resonance bandwidth $2\Delta\omega_{0,k}$ from the intersections with the circle segments $|X_k|^n = R_{in}^k$. The reflection for $f \gg f_k$ or $f \ll f_k$ is not at the short circuit point $-1$. It is shifted along the circumference of the Smith chart. The reference plane is automatically shifted to obtain a pure parallel resonance. Fig. 7 shows an example for a well-isolated resonance at $f_k = 515.99$ MHz.

The determined values are $K_{1,k} = 0.89$ which is near to critical coupling and $Q_{0,k} = 146 400$. According to [11], a nearby resonance of the empty chamber at 516 MHz (with the integers 5, 4, 1) has theoretical values $Q_{0,k} = 196 000$ (TM$_z$ mode) and $Q_{0,k} = 162 000$ (TE$_z$ mode).

All measurements regarding electrical mode stirring are made in frequency bands with a width between 1 and 100 MHz.

Fig. 7. Locus of the input resistance for a resonance at $f_k = 515.99$ MHz in the Smith chart. The ratio $R_{in}^k / Z_L$ at resonance determines the coupling factor; the unloaded quality factor $Q_{0,k}$ is determined from the bandwidth $2\Delta\omega_{0,k}$ between the locus and the circle segments $|X_k|^n = R_{in}^k$.

![Fig. 7. Locus of the input resistance for a resonance at $f_k = 515.99$ MHz in the Smith chart. The ratio $R_{in}^k / Z_L$ at resonance determines the coupling factor; the unloaded quality factor $Q_{0,k}$ is determined from the bandwidth $2\Delta\omega_{0,k}$ between the locus and the circle segments $|X_k|^n = R_{in}^k$.](image)

All measurements yield basically the same results. Therefore, we only present results for a narrow 1-MHz band centered at 2 GHz. In this case, the resonances are distinguishable. Fig. 8 presents the magnitudes $|s_{11}|$, $|s_{22}|$, and $|s_{21}|$ of the two-port between antennas 1 and 2 in Fig. 4.

![Fig. 8. Magnitudes $|s_{11}|$, $|s_{22}|$, and $|s_{21}|$ of the two-port between antennas 1 and 2 in Fig. 4.](image)

All measurements yield basically the same results. Therefore, we only present results for a narrow 1-MHz band centered at 2 GHz. In this case, the resonances are distinguishable. Fig. 8 presents the magnitudes $|s_{11}|$, $|s_{22}|$, and $|s_{21}|$ of the two-port with the antennas 1 and 2 (see Fig. 4). Antenna 2 is loaded by $Z_L$, and antennas 3 and 4 are open circuited. In this case, the loaded quality factors $Q_{L,k}$ are given by

$$\frac{1}{Q_{L,k}} = \frac{1}{Q_{0,k}} + \frac{1}{Q_{ext1,k}} + \frac{1}{Q_{ext2,k}}. \quad (13)$$

(In case of a critical coupling of both antennas, we have $Q_{L,k} = Q_{0,k}/3$.)

We observe resonances with low reflection coefficients $|s_{11}|$, $|s_{22}|$. These resonances are almost critically coupled; we have $K_{1,k} \approx K_{2,k} \approx 1$. For low simultaneous values of $|s_{11}|$ and $|s_{22}|$, the transmission coefficient $|s_{21}|$ reaches relatively high values; both antennas are well coupled. On the other hand, many other resonances are under-/overcoupled with subsequently low values of $|s_{21}|$. These differences between an almost critical coupling (which is optimal) and a strong under-/overcoupling occurs within a few 10 kHz frequency separation. Each resonance frequency $f_k$ of the empty chamber is determined by three integers $m$, $n$, and $p$, according to [9]–[11]

$$f_k = \frac{c_0}{2} \sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{c}\right)^2} \quad c_0 = \frac{1}{\sqrt{\varepsilon_0\mu_0}}. \quad (14)$$

Neighboring resonances usually correspond to widely different sets $m$, $n$, and $p$. Correspondingly, they have significantly different field distributions. Therefore, the coupling coefficients can also vary significantly. For example, the three resonant frequencies of the empty chamber nearest to 2000 MHz are 1999.83/1999.90/2000.13 MHz. They are determined by the sets $m,n,p = (14,21,2)/(1,17,29)/(17,17,11)$.

The antennas break the degeneracy between the TE$_z$- and TM$_z$-modes that exists for $m,n,p \geq 1$. A perturbation approach [9]–[11] that employs the modes with fields $H_0$ and $E_0$ of the empty chamber yields a shift of the resonance frequencies

$$f - f_k \approx f_k \left\{ \iint_{\text{antennas}} (\mu_0 |\vec{H}_0|^2 - \varepsilon_0 |\vec{E}_0|^2) dv \right\} \left\{ \iint_{\text{chamber}} (\mu_0 |\vec{H}_0|^2 + \varepsilon_0 |\vec{E}_0|^2) dv \right\}. \quad (15)$$
In our case, (15) is only a crude approximation, because (15) is well applicable only for shallow wall deformations. Calculations for the three resonance frequencies given earlier show a typical frequency shift of the four antennas for the TE\_z- and TM\_z-modes in the range -1.3 MHz to +0.5 MHz. It is practically impossible to assign integers (m, n, p) to measured resonances.

We qualitatively show the significant change of the coupling factor by using the known formalism [10], [11] of aperture coupling to cavities. The aperture is shown in the cross-sectional Fig. 5 by the dotted circle segments. The fundamental TEM transmission line wave of the antenna has a polar electrical field in the aperture

\[ E_{\text{ant}} = \frac{Z_0}{2\pi \sqrt{Z_L}} \frac{1}{l \sin \vartheta} l = \frac{h}{\cos \vartheta_0} \]  

which is normalized to a unit forward power. The unperturbed mode fields are normalized to a unit electrical or magnetic energy of the resonant fields. Numerical time-domain calculations show strong field distortions in the vicinity of the antennas. Therefore, unperturbed fields are again only a crude approximation and a complete derivation is not presented. The coupling coefficients for our choice of the aperture are then proportional to the surface integral [10], [11]

\[ C = \left| \int_{\vartheta_0}^{\pi/2} \int_{0}^{2\pi} E_{\vartheta}^{\text{reso}} H_{\varphi,k}^{\text{reso}} \sin \vartheta d\varphi d\vartheta \right|^2 \]  

where \( H_{\varphi,k}^{\text{reso}} \) is expressed by the Cartesian components of the magnetic mode fields. The calculations of C for antenna 1 show for neighboring resonances \( f_k \) a variation of C by more than an order of magnitude, with large differences between the TE\_z- and TM\_z-modes. We expect that many resonances cannot be detected experimentally.

Different techniques are feasible for a variable reactive loading.

1) Electromechanical multiport switches with short or open-circuited transmission line elements have low losses.
2) Alternatively, pin-diode switches can be used.
3) Field-effect transistor (FET) switches usually have higher losses.
4) Varactor diodes for higher power are usually not available.

Here, we use variable coaxial short circuit plungers because of the possibility to select appropriate positions. Each plunger is set to four positions. In our case of three reactively loaded antennas and four positions per plunger, we have 64 combinations of reactive loading the chamber. (This approximately corresponds to the number of rotation positions for mechanical mode stirrers.)

Fig. 9(a) shows \(|r_{\text{in}}|\) in the frequency range 1999.5–2000.5 MHz (compare Fig. 8) for all the 64 combinations. We recognize the considerable amount of mode stirring. There are, however, frequency intervals, where \(|r_{\text{in}}|\) is rather high. In these intervals, the antennas are only weakly coupled to the resonances. In these measurements, the unloaded quality factors \(Q_{0,k}\) are around 100 000. In further measurements, the quality factors \(Q_{0,k}\) are reduced to about 5000 by placing small absorbers in the chamber. Similar results as in Fig. 9(a) are obtained. The resonances are, of course, broader in this case.

A first indication of field isotropy and homogeneity—although no proof—is an antenna input reflection \(|r_{\text{in}}|\) close to free-space radiation [13]. Therefore, we depict in Fig. 9(b) the mean value \(|\langle r_{\text{in}} \rangle|\) of the input reflection. (Full line) \(Q_{0,k} \approx 100 000\). (Dashed line) \(Q_{0,k} \approx 5000\).

The measurements in other frequency ranges essentially confirm the aforementioned results. As discussed in connection with Fig. 8, the origin of this behavior is the significant change of the field distributions with resonant frequency. Therefore, one should investigate optimal antenna positions and orientations or should design antenna systems, which are more insensitive to changes of the field distribution.

As a first basic statistical investigation, we calculate the scatter plot for the measured transmission coefficient \(s_{51}\) between the transmitting antenna 1 and the receiving antenna 5, which is placed in the working volume of the chamber. The measurements are made with 64 plunger combinations and an empty chamber. The results are depicted in Fig. 10 for a frequency of \(f = 1996.96\) MHz.

The significant changes of the inphase and quadrature component of \(s_{51}\) indicate that the plunger combinations are correlated only weakly [5].

The weak correlation, respectively, good independence of the plunger positions is confirmed by a low magnitude of the complex autocorrelation function of the transmission coefficient \(s_{51}\). For independence, the International Electrotechnical Commission (IEC) standard 61000-4-21:2003 calls for an
N = plunger combinations and \( f \) as independent. This applies to all analyzed frequencies.

The maximum autocorrelation coefficient for an arbitrary offset is less than 0.2; thus, the plunger positions can be assumed as independent. This applies to all analyzed frequencies.

**IV. Conclusion**

The concept of electrical mode stirring in reverberating chambers for EMC by reactively loaded antennas is demonstrated theoretically and experimentally. A network model of the chamber with measurable parameters is used to explain the operation principle and to derive basic design criteria. Extensive experimental investigations in the gigahertz range for a chamber of 10.3 m\(^3\) volume equipped with conical antennas demonstrate a considerable mode stirring by reactively loaded antennas and show the applicability of this novel approach. In further work we will investigate antennas and antenna configurations for an optimal coupling of the antennas to the resonant fields. Further investigations of the field homogeneity and isotropy and further investigations of the field statistics will be performed.

**REFERENCES**


**Edgar Voges** was born in Braunschweig, Germany, in 1941. He received the Dipl.-Phys. and Dr.-Ing. degrees from the Technische Universität Braunschweig, Braunschweig, Germany, in 1967 and 1970, respectively.

From 1974 to 1977, he was a Professor in the Department of Electrical Engineering, University of Dortmund, Dortmund, Germany, where he has been the Head of the High Frequency Institute, Department of Electrical Engineering and Information Technology since 1982. From 1977 to 1982, he was a Professor of communication engineering at the Fern Universität, Hagen, Germany. His current research interests include optical communication, integrated optics, and microwave circuits.

**Tobias Eisenburger** was born in Witten, Germany, in 1980. He received the Dipl.-Ing. degree from Ruhr-Universität Bochum, Bochum, Germany, in 2006.

He is currently with the Department of Electrical Engineering and Information Technology, University of Dortmund, Dortmund, Germany. His current research interests include electromagnetic compatibility and antennas.