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Investigation of High-Q Resonators Embedded in Low Temperature Cofired Ceramic for Millimetre-Wave Oscillators

by

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Abstract

This thesis proposes a novel approach to realizing low cost, miniaturized, low phase noise, medium power, free running millimeter wave oscillators by embedding a high Q laminated waveguide (LWG) resonator in the same low temperature cofired ceramic (LTCC) package that houses the monolithic oscillator active circuit. The proposed technique overcomes the drawbacks of hybrid waveguide/dielectric resonator oscillators and fully monolithic integrated oscillators.

As a major step toward achieving this result, a comprehensive study of LTCC resonator configurations is carried out with the aim of achieving high Q values at millimeter wave frequencies. Rectangular and cylindrical resonator geometries with various feeding mechanisms (probe, loop, coplanar line) and filling materials (air, dielectric) are analytically and experimentally investigated. A design strategy for obtaining high Q factors using a number of loss-reduction and feeding techniques is proposed. A few optimum designs are then implemented, yielding unloaded Q values of 700 to 1100 at Ka-band, the highest values ever reported for a standard LTCC process. Excellent agreement was obtained between EM modeling and measurements.

The final thesis concept is demonstrated through two oscillators; one for the 24 GHz ISM and the second for the 29 GHz LMDS band. An active monolithic microwave integrated circuit (MMIC) chip is integrated with the high-Q LTCC resonators to realize the proposed high performance oscillators. For the 29 GHz oscillator, the measured output power is 11.5 dBm at 28.9 GHz with dc-conversion efficiency of 6.5%. The free running phase noise is -101.5 dBc/Hz at 100 KHz offset, which is among the best results ever published for a free running K/Ka-band oscillator. This represents the first reported instance of a low phase noise millimeter-wave oscillator utilizing a high Q resonator embedded in the same LTCC package.

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List of Acronyms and Abbreviations

ADS	Analog Design System
BGA	ball grid array
CAD	Computer Aided Design
CB	common base
CE	common emitter
CPS	coplanar strip line
CPW	coplanar waveguide
CST	computer systems technology
CTE	coefficient of thermal expansion
DR	dielectric resonator
DRO	dielectric resonator oscillator
DUT	device under test
EBG	electromagnetic band-gap
EC	electromagnetic crystals
EM	electromagnetic
FDTD	finite-difference-time-domain
FEM	finite element method
FFT	fast fourier transform
FIT	finite integration technique
FM	Frequency modulation
GaAs	gallium arsenide
GHz	gigahertz
GPIO	General purpose interface bus
HBT	heterojunction bipolar transistor
HEMT	High electron mobility transistor
ISM	Industrial, scientific medical band
Ka-band	26-40 GHz
K-band	18-26 GHz
LMDS	Local multi-point distribution service
LTCC	low temperature cofired ceramics
LTCC-M	LTCC on metal
LTI	Linear time invariant

LTV	Linear time variant
LWG	laminated wave guide
MCM_L	laminated multi chip modules
MEMS	Microelectromechanical systems
MESFET	Metal semiconductor field effect transistor
MIC	microwave integrated circuit
MMIC	monolithic microwave integrated circuit
MMW	millimeter wave
MW	microwave
OSLT	open, short, load, through
PBA	perfect boundary approximation
PBG	photonic band-gap
PHEMT	Pseudomorphic HEMT
PLL	phase-locked loop
PS	Power supply
Q	quality factor
R	resistance
RF	radio frequency
RWG	rectangular waveguide
SiGe	silicon germanium technology
SIW	substrate-integrated waveguide
SL	slot-line
SSB	single-sideband
TEM	transverse electromagnetic
TL	transmission line
VCO	voltage controlled oscillator
VSWR	voltage standing wave ratio
VWG	via wave guide
YIG	yttrium-iron garnet
μ BGA	micro ball grid array

Glossary

H_ϕ	phi-component of H field in a cylindrical co-ordinate	I_{DS}	Drain current in a FET
$E_{o_{max}}$	maximum electric field value at first resonance	j	Aquare root of (-1)
$H_{o_{max}}$	maximum electric field value at first resonance	J_n	Nth-order Bessel function of 1 st order
$\bar{E}(x)$	Normalized value of electric field	k	Boltzmann's constant = 1.38×10^{-23} J/K
$E(x)$	Electric field as a function of x-direction	K(S)	Stability factor
f_{o1}	resonance frequency of first mode	L	length, inductance
κ	coupling coefficient, permeability tensor variable	l	overall resonator thickness
a	radius of a cylindrical resonators or width of rectangular resonators	L_{air}	Number of layers used in the air cavity
A	side cavity side length	l_{diel}	Number of layers of the dielectric region
A(s)	Amplifier gain transfer function	L_{fb}	Length of Feed back stub
a_{eff}	Effective radius of a cylindrical resonator	l_{total}	total number of layers
b	height	L_{total}	total thickness of the cylindrical resonator
B	magnetic field	$\mu(S)$	geometrically derived stability parameter
B(s)	frequency selective network transfer function	n(t)	noise source
B_{IN}	input susceptance	p	via pitch
C	capacitance, auxiliary variable	phi	diameter
C_{gd}	transistor gate-drain capacitance	P_{lc}	average conductor power loss in a metallic resonator
C_{RF}	DC blocking capacitance	P_{ld}	power loss due to the dielectric only
d	air cavity depth ratio in a laminated resonator	P_o	output power
E	electric field intensity	P_s	amplifier input power level
f	frequency	Q	quality factor
F	active device thermal noise factor	Q_{o1}	Unloaded Q of first mode
f_c	flicker-noise corner frequency	Q_c	conductor quality factor
f_m	frequency offset from carrier	Q_c	conductor quality factor
f_o	resonant frequency	Q_d	dielectric quality factor

f_o	oscillator fundamental frequency	Γ_d	device reflection coefficient
H	magnetic field, auxiliary variable	Q_{ext}	external quality factor
H_t	peak value of the tangential magnetic feild	Q_L	loaded quality factor
Q_o	unloaded quality factor	Q_l	Loaded quality factor of the resonator
Q_r	radiation Q factor	T_C	Critical temperature for superconductors
Q_r	radiation Q factor	Γ_c	circuit reflection coefficient
R_d	surface resistance		
R_o	Characteristic resistance		
R_s	surface resistance of resonators		
S	S-parameters		
S_{11}	input reflection coefficient		
t	layer thickness		
T	temperature °K		
V_d	drain bias voltages		
V_{gs}	gate bias voltage		
w	energy		
W	width		
W_e	electric energy		
W_m	magnetic energy		
X_{IN}	input reactance		
Y_c	Circuit admittance		
Y_d	Device admittance		
z, r and ϕ	Cylindrical co-ordinate		
Z_{IN}	Input impedance		
Z_o	Cherecteristic impedance		
δ	skin depth		
ϵ	permittivity of dielectric substrate		
ϵ_{eff}	Effective dielectric constant		
ϵ_r	relative permittivity constant		
η	Intrinsic impedance		
μ	permeability of dielectric substrate		
σ	conductivity		
$\phi(t)$	phase noise function		
ω_o	resonant frequency		

CHAPTER 1 Introduction

1.1 Motivation

The lower GHz spectrum has been saturated by wireless communication systems used mainly for voice and data transmission through cellular networks [1]. Today's emerging applications call for larger bandwidths to accommodate the transfer of multimedia information to roaming users. As a result, new transceivers are becoming the critical components needed for such expanding applications, with their carrier frequencies being pushed into the tens of gigahertz. Microwave (MW) and Millimeter wave (MMW) frequencies have been used for a few decades mostly for military and space applications (radar, electronic warfare and satellite communications). Recently, many more commercial applications have appeared, such as mobile telecommunications, smart automotive anti-collision radars, point-to-point digital radio, security sensors, and biomedical applications [2], [3]. Because of this accelerating growth of MMW technology, the market is in critical need of low cost, low volume, high performance transceivers. One important element of any transceiver is the oscillator, and its design has become challenging due to carrier phase noise requirements for bit error rate enhancement. In this context, an overriding concern of the work herein is the development of low cost, miniaturized, low phase noise oscillators.

Two technology extremes currently exist. The first is to embed all the system components on a chip using monolithic microwave integrated circuit (MMIC) technology. This reduces the cost and the size, as shown in Figure 1.1, but still has

relatively low performance due to the limitation of using planar resonator circuits, which have poor quality factors [4]-[6]. The other extreme is to use cryogenically-cooled hybrid waveguide assembly as shown in Figure 1.2 [7]. This technique can yield the highest phase noise performance, but it suffers from the large cost and the bulky volume of the resonators and other waveguide components [8], [9], [10]. Between these two extremes lies fertile ground for technology development that will lead to the desired levels of performance

Figure 1.1 Micrograph of a 25 GHz VCO MMIC chip using SiGe Technology [6]

As an example, the hybrid assembly of a high quality dielectric disk and a planar active circuit yields dielectric-resonator (DR) oscillators with excellent performance in terms of close-to-carrier phase noise [11], [12]. But, these need careful post-fabrication tuning to position the dielectric resonator on the substrate. The demanding factors of cost, size, reliability, and repeatability for emerging systems still require another solution.

Figure 1.2 Photograph of a Cryogenic low phase noise Saphire 23 GHz oscillator [7]

The rapidly growing field of wireless communications is providing new opportunities to develop novel structures that will enhance or even replace existing circuits and systems. The increasing popularity of millimeter wave applications has created many challenges for interconnect and packaging technologies. Cofired ceramics have found emerging acceptance in the packaging of microwave integrated circuits due to the ability of fabricating planar microwave structures and embedding passive elements within the laminations of the package. Low temperature cofired ceramics (LTCC) combine the beneficial electrical properties of ceramics at mm-wave frequencies with a precise control of the dielectric thickness. Therefore, LTCC would be a promising technology for embedding the resonator with the MMIC chips [13]-[15], so as to produce small size, low cost and potentially high performance oscillators.

The implementation of a resonator in LTCC using rows of vias as sidewalls was initially presented in [16]. A number of drawbacks resulted in a low Q (only ~ 150) at 5.8 GHz. Despite the importance of achieving compact, mass producible high-Q resonators

at higher frequencies, very little has been published in this area and, to the author's knowledge, nothing in LTCC

1.2 Thesis Objectives

The overall goal of this work is to develop high-performance K/Ka band oscillators in a novel LTCC packaging environment. Several barriers must be overcome to achieve such a result, and these form the specific objectives of this thesis

The first objective is to investigate high quality resonators using several configurations of the LTCC environment. A significant challenge is to obtain suitable resonator structures, which obey typical LTCC design rules and at the same time provide high Q in the MMW frequency range.

A resonator is of no practical use unless it can be connected to external circuitry and excited correctly. Thus, the second objective of this thesis is to investigate appropriate feeding mechanisms that allow sufficient control of the excited modes inside the LTCC resonators.

Since some of the resonators developed herein are completely novel, the third thesis objective is to perform a field analysis as well as structural parametric analysis in order to fully understand the operational capabilities and limits of the candidate structures.

The fourth and final objective of this thesis is to demonstrate the feasibility of employing the proposed LTCC high-Q resonators to achieve high-performance oscillators through the design, optimization, fabrication and measurement of prototypes.

1.3 Thesis Organization

This thesis consists of five additional chapters. Chapter 2 establishes the state of the art in resonator technology at microwave and millimeter-wave frequencies, in which the concept of resonance and the definition of the performance parameters are introduced. An exhaustive overview of the types of resonators follows, along with various comparisons between published microwave and millimeter-wave resonators.

Chapter 3 is dedicated to the investigation of high-Q LTCC resonance structures at microwave and millimeter-wave frequencies, with emphasis placed on techniques that maximize the quality factor and are compatible with standard LTCC processing. Resonant structures are divided into rectangular and cylindrical categories. Excitation methods are also developed and adapted to LTCC resonators. Then, novel excitation using the centre strip of planar lines is proposed and found to be very promising. Finally, some new LTCC resonators are designed, modeled, optimized and verified experimentally.

Chapter 4 presents an analytical investigation of the problem of cylindrical multilayered dielectric resonators with rectangular air cavities inside. Different modes that can be excited inside this structure are studied along with the effect of the key parameters of the structure. A filling factor is defined to relate the ratio of the air volume to the whole structure volume and to be used in different case studies. Also, the effect of the excitation method and the position of the feeding element on both frequency and Q-factors are presented. Then, a baseline-optimized design of a cylindrical laminated waveguide resonator is carried out and tested.

In Chapter 5, the theory of microwave and millimeter wave oscillators and a review of the recent literature is given. A brief overview of oscillator phase noise models is also presented. Then, a proposed low phase noise GaAs MMIC oscillator is designed, optimized and integrated within selected high Q LTCC resonators of chapter 3. Prototypes have been manufactured and their measured results are provided therein.

The final chapter contains the conclusions, lists the contributions of this thesis, and suggests future work to enhance the performance of the proposed MMW resonators and oscillators. Four appendices follow: Appendix A provides specification data sheets for the LTCC material employed along with its measured parameters, Appendix B describes the GaAs foundry process which is used to implement MMIC chips for this work, Appendix C explains the employed method to measure Q from the resonator S-parameters, Appendix D contains the characteristics of the probes and cables used in the measurements to help de-embed the spectrum analyzer results of Chapter 5.

CHAPTER 2 An Overview Of Microwave Resonators

2.1 Introduction

Resonant structures are extensively used as network elements in the realization of microwave oscillators and filter circuits. As the frequency increases, lumped elements cannot be used to build high-Q resonant circuits. Depending on the application, their realization varies widely. Therefore, in this chapter many resonant structures will be discussed, as well as their excitation techniques, analysis and design methods. Special attention will be given to waveguide cavity resonators, as these will be used in the following chapters. Before describing the different structures, some basic parameters of microwave resonators will be defined. At the end of this thesis, a method to extract the resonator parameters from Scattering parameters is explained.

2.2 Resonator Parameters

Microwave energy can be stored in resonator circuit consisting of either lumped elements, which are small compared with a wavelength, or of distributed elements, which are of a size comparable with a wavelength. Whatever the type, the following basic resonator parameters apply [17].

2.2.1 Resonant frequency and equivalent models

Resonant frequency is the frequency at which the electric and magnetic stored energies are equal; i.e. the input impedance of the circuit is purely a real value. It is more convenient to present the resonant frequency definition through the concept of lumped

element models. Physical unloaded resonators may be approximated by lumped element series or parallel RLC circuits in the vicinity of their resonant frequency, as shown in Figure 2.1. The input reactance X_{IN} reflects the ideal behavior of the resonator, while the resistance R represents the loss mechanisms inside that resonator. It is known that a parallel resonance presents its highest impedance magnitude to the input port at the resonant frequency f_o , while a series resonance presents its lowest impedance magnitude at f_o . A series RLC model has a constant resistance, while a parallel cavity displays a constant conductance. Figure 2.2 shows how the input reactance behaves near resonance. In the case of series resonance, the input reactance passes from negative to positive as the frequency increases beyond resonance and it varies slowly with frequency:

$$X_{IN} = \frac{-1}{\omega C} + \omega L \quad (2.1)$$

While at resonant frequency ω_o , $X_{in}=0$ and thus

$$\omega_o = \frac{1}{\sqrt{LC}} \quad (2.2)$$

In the case of a parallel resonance model,

$$B_{IN} = \omega C - \frac{1}{\omega L} = \frac{\omega^2 LC - 1}{\omega L} \quad (2.3)$$

$$X_{IN} = \frac{\omega L}{1 - \omega^2 LC} \quad (2.4)$$

So, the denominator goes from positive to negative as the frequency increases beyond resonance. The denominator is exactly zero at resonance, yielding infinite reactance, which is clear from the asymptotic behavior of the reactance near resonance,

as shown in Figure 2.2 (b). Therefore, the resonance frequency of the simplest lumped-element model with parallel or series components is always given by:

$$f_o = \frac{1}{2\pi\sqrt{LC}} \quad (2.5)$$

Figure 2.1. Simplest series and parallel lumped-element models for an unloaded resonator in the vicinity of its resonant frequency.

Figure 2.2. The input reactance behavior near resonance for series and parallel resonators.

One might expect the Q of the series case to be very small, given the very slow change in X_{IN} with frequency. However, a series resonance presents an impedance minimum at resonance, so the slow-changing X_{IN} actually represents a significant reactance change relative to the low impedance value. On the other hand, parallel resonance presents an impedance maximum, so it is not surprising that the X_{IN} variation must be great to cause the input impedance to change.

From another point of view, resonators may be divided into one-port (reflection type) and two-port (transmission) types. Therefore, in the one-port case, the resonance frequency can be defined as the frequency at which minimum reflection occurs; i.e. the energy absorbed in the resonator is at its maximum. On the other hand, the resonance frequency of the 2-port is the frequency at which maximum transmission occurs; i.e. the energy absorbed in the resonator is minimum.

2.2.2 Quality factor Q:

The main characteristic of the resonator is the quality factor. In terms of energy ratio it is defined as [18]:

$$Q = \omega \cdot \frac{(\text{time-averaged energy stored in the resonator})}{\text{energy loss per second in the resonator}} \quad (2.6)$$

The amplitude of the stored energy will decay because of the resonator circuit's electrical resistance. The rate of decay depends on the resonator's quality factor Q . The lower the Q , the greater the rate of decay. High- Q resonators can be used to build very stable, low noise oscillators. Low- Q resonators are employed in broadband tuned oscillators.

Figure 2.3 shows the frequency response of the input impedance of the ideal lumped-element models of Figure 2.1. It is clear that the higher the Q of series resonators, the lower its minimum Z_{IN} , while the higher the Q of parallel resonators, the larger its Z_{IN} . Figure 2.4 shows the corresponding impedance loci on a Smith Chart.

For distributed resonator circuits, the total energy stored in the resonator is obtained by integrating the energy density over the resonator volume. The electric and magnetic energies are given by [19]:

$$W_e = \frac{\varepsilon}{2} \int_v |E|^2 dv \quad (2.7)$$

$$W_m = \frac{\mu}{2} \int_v |H|^2 dv \quad (2.8)$$

At resonance both energies should be equal, i.e.

$$W_e = W_m = W$$

On the other hand, the sources of losses inside a resonator may be due to conductor, dielectric or radiation losses. The average conductor power loss in a metallic resonator is obtained by integrating the power density over the inner metal surface of the resonator.

$$P_{lc} = \frac{R_s}{2} \int_s |H_t|^2 da \quad (2.9)$$

where: R_s is the surface resistance of the resonator.

So, Q_c due to the conductor loss, assuming lossless dielectric filling, can be obtained as follows

$$Q_c = \frac{\omega W}{P_{lc}} = \frac{\omega \mu \int_v |H|^2 dv}{R_s \int_s |H_t|^2 da} \quad (2.10)$$

where: H_t is the peak value of the tangential magnetic intensity.

Figure 2.3. The input impedance versus frequency for series and parallel resonators.

Figure 2.4. Smith Chart input impedance loci of undercoupled resonators, (a) parallel and (b) series.

Q_d is the Q when a lossy dielectric medium is present but the walls are perfectly conducting assuming no radiation;

$$Q_d = \frac{\omega W}{P_{ld}} \quad (2.11)$$

where: P_{ld} is the power loss due to the dielectric only.

The radiation Q factor, denoted Q_r , is defined as [19]:

$$Q_r = \frac{\omega W}{P_{rad}} \quad (2.12)$$

So, the resultant unloaded Q , denoted Q_o of a resonant cavity is given by

$$\frac{1}{Q_o} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_r} \quad (2.13)$$

where Q_r is the quality factor due to radiation losses only. Q_o is called unloaded Q because the dissipated power is calculated inside the structure only without including any coupling to an external load. But, in a practical situation, the resonant circuit is coupled to a load that also dissipates power. If the power dissipation is calculated inside the external load only, the quality factor will be denoted Q_{ext} ; i.e. Q_{ext} is the external quality factor for a lossless resonator in the presence of the load. On the other hand, if the Q is due to both load coupling and resonator losses it will be denoted as loaded quality factor Q_L . The loaded quality factor Q_L is given by:

$$\frac{1}{Q_L} = \frac{1}{Q_o} + \frac{1}{Q_{ext}} \quad (2.14)$$

One of the objectives of this research is to obtain a cavity-like resonator with high Q_o . To obtain the highest Q_o , the cavity should be made as large as possible to increase the volume to conductor ratio, but without introducing spurious modes. Also, if the coupling between the cavity and its input is sufficiently weak, Q_L will be high.

2.2.3 Mode interference

It is well known that a given cavity has an infinite number of resonant frequencies. The complete set of normal modes can be calculated from solving the wave equation. The lowest frequency of the set is called the *Dominant Mode*. Mode interference is governed by the size of the cavity in relation with wavelength λ . At best: it should be decided which particular mode of the infinite set will be excited. At worst: the resonance frequencies may be too close together to distinguish one frequency from another. Fortunately, not all of the possible interfering modes will cause difficulty in practice because of the *mode suppression possibility*; if the field distribution is known, input and output coupling systems can prevent undesirable modes. In addition, slots or resistive wires may be adjusted to attenuate the spurious modes without disturbing the desired mode [20]. Another objective of this thesis is to propose some high-Q resonators, study their modes, and try to suppress the unwanted modes in each if there is any.

2.2.4 Coupling Mechanisms:

Coupling structures provide a means of coupling energy into and out of the cavity. It may be accomplished by an inductive current loop or hole in the cavity wall, or by means of capacitive electric probes. Figure 2.5 shows different coupling mechanisms for metallic cavity resonators as well as a simplified model for each.

When loop coupling is used, the orientation of the loop should be in a plane perpendicular to the H field of the excited mode. For maximum coupling, usually the loop is placed in a region of maximum magnetic field intensity. The size of the loop also determines the degree of coupling.

For hole coupling, attention must be paid to the orientation of the external waveguide with respect to the resonator so that the desired mode is coupled between the two. The size of the coupling aperture determines the degree of coupling.

It is obvious from the equivalent circuit of Figure 2.5 that the coupling affects the resonance frequency and the Q of the system. To simplify the understanding of the coupling, a conventional lumped parameter RLC parallel network coupled by means of an ideal N:1 transformer is shown in Figure 2.6. A coupling coefficient κ is defined to be the ratio of the power dissipated in the external circuit to the power dissipated within the shielded cavity.

$$\kappa = \frac{P_{ext}}{P_o} = \frac{Q_o}{Q_{ext}} = \frac{R_o}{N^2 Z_o} \quad (2.15)$$

Where, the resistance R_o accounts for the power loss in the resonator circuit. This parameter is a good measure for the amount of coupling and can lead to a classification of the resonator coupling into three types [20]:

- Critical coupling, if the resonator is matched to the generator; $\kappa = 1$. In this case the power dissipated in the external circuit equals the power dissipated in the unloaded cavity.
- Over coupling, if the resonator terminals are at a voltage maximum, $\kappa > 1$ and

$$\kappa = VSWR \quad (2.16)$$

where VSWR is the voltage standing wave ratio. In this case the resonator could be modeled by the series lumped-element model of Figure 2.1(b).

- Under coupling, if the resonator terminals are at a voltage minimum; $\kappa < 1$

$$\kappa = \frac{1}{VSWR} \quad (2.17)$$

In this case the resonator could be modeled by the parallel lumped-element circuit of Figure 2.1(a).

(a) Iris coupling to a rectangular cavity resonator.

(b) Loop coupling of a cavity resonator and the equivalent circuit of both.

(c) Probe coupling of a cavity resonator and the equivalent circuit of both.

Figure 2.5. Examples of coupling mechanisms of typical cavities [19]

Figure 2.6. The equivalent circuit of a resonant cavity with one coupling system (coupled to a TL) [20]

2.3 Types of resonators

Future high-Q resonators will be dictated by a shift in operating frequency of communications systems towards shorter wavelengths, mainly the millimeter-wave regime, to avoid the frequency congestion at lower frequencies. The essential question in modern applications is, whether or not the resonator structure is mass producible. Therefore, in this section many types of resonators are discussed with their advantages, disadvantages and suitability for specific applications. Some other types like crystal resonators and acoustic wave (AW) resonators are not mentioned because they are only used at very low microwave frequencies.

2.3.1 Planar resonators

Planar resonators are used in MMICs such as oscillators but also in directional couplers, circulators and filters. A microstrip resonator is a planar conductor, which is deposited on a dielectric substrate. It may take several shapes, depending on the applications. Common, microstrip shapes are rectangular, circular disk, circular ring, triangular and rhombus resonators. The amplitude of the current that flows on the structure becomes significant when the signal frequency is close to a resonance. Resonance starts to occur when its size is about a half-wavelength [21].

Planar resonators are easier to design than three-dimensional resonators. On the other hand, they offer much lower Q values. As an example, microstrip rectangular, circular and ring resonators will be introduced. Although such planar resonators are commonly used, they are still limited in Q (less than 300). The structure of a rectangular resonator is shown in Figure 2.7. An accurate Q factor and resonance frequency determination are given by Itoh [22]:

$$f_0 = \frac{\sqrt{\left(\frac{1}{2w}\right)^2 + \left(\frac{1}{2l}\right)^2}}{2\sqrt{\epsilon\mu}} \quad (2.18)$$

where: ϵ = permittivity of the microstrip substrate

μ = permeability of the microstrip substrate

$$Q = \frac{\omega_0 d \mu}{4R_s} \quad (2.19)$$

where:

$$R_s = \text{surface resistance} = \sqrt{\frac{\pi f \mu}{\sigma}}$$

σ = conductivity of the microstrip conductor.

Wolff and Knoppik [23] give a method to calculate resonant frequency within 1 to 2%. While an improvement to 0.5% has been reported by Verma and Rostany [24].

Figure 2.7. Microstrip Rectangular Resonators [22]

A circular resonator with radius a is also given by Wolff and Knoppik [23]. They calculated f_o to within approximately 1%. Bahl and Bhartia [25] reviewed other calculation methods. The closed and open microstrip ring resonators are shown in Figure 2.8. Wu and Rosenbaum [26] present a mode chart for the (closed) ring resonator. Khilla [27] presents a method to calculate f_o to within 1.5%.

$$f_o = \frac{3.832}{2\pi a \sqrt{\mu\epsilon}} \quad (2.20)$$

$$Q = \frac{\omega_o d\mu}{2R_w} \quad (2.21)$$

Tripathi and Wolff [28] present closed-form solutions for both the open and closed ring resonators [29].

Figure 2.8. Closed and open ring resonators [23].

2.3.2 Dielectric Resonators

The dielectric resonator is made of a low-loss, temperature-stable, high-permittivity, and high- Q ceramic material in a regular geometrical form. It resonates in various modes at frequencies determined by its dimensions and shielding conditions. Because of its small size, low price, and easy integrability in hybrid microwave integrated circuits (HMICs), its application in active and passive microwave circuits has been rapidly increasing [30],[31]. The shape of a dielectric resonator is usually a solid cylinder and its resonant frequency is determined by its dimensions and surroundings. Dielectric resonators are used primarily in the 3- to 30-GHz frequency range. Below 3 GHz the physical size of the resonator becomes prohibitively large for most applications.

Over the past two decades, the DRO (Dielectric resonator oscillator) became the oscillator of choice for fixed frequency applications requiring high stability and low phase noise. A high- Q temperature stable dielectric resonator is used as the key element in the design of oscillator configurations shown in Figure 2.9.

Finally, all DRs are hardly amenable to mass fabrication due to their frequency-sensitivity, although they can offer high Q 's (up to 20,000). In addition, they are individually tuned to specifications, which is considered to be their main disadvantage.

Figure 2.9. Standard DRO Design Topologies [30],[31]

2.3.3 Photonic Band Gap Resonators

Photonic band-gap (PBG) are periodic dielectric and/or metallic structures that, when designed and implemented correctly, can improve the performance of mm-wave devices or antennas. PBG, sometimes referred to as electromagnetic band-gap (EBG), electromagnetic crystals (EC) or photonic crystals, are inhomogeneous structures. They are usually composed of periodic regions of a certain material immersed in a homogeneous background of another material with a different permittivity. In solid state theory, semiconductors allow electron conduction without scattering only for electrons that have enough energy often termed as band-gaps. Electromagnetic propagation in PBG materials is analogous to electron wave propagation in semiconductors, though different propagation mechanisms are involved. McCall et. al. [32] were the first to calculate and measure microwave propagation and reflection in a two-dimensional (2D) array of low-loss high dielectric constant cylinders. Reviews of PBG research can be found in special issues of [33], [34] as well as a recent book [35].

A one-dimensional 1D-PBG resonator can exist in coplanar waveguide (CPW), coplanar strip line (CPS) and slot-line (SL) as shown in Figure 2.10 [36]. The reported unloaded Q for the implemented SL resonator was 299 at 9.3 GHz, using RT/Duroid 6010.5 substrate. The idea of building this kind of PBG resonator is to create a defect in

the structure with length L_R greater than the cell periodic distance d as shown in Figure 2.10. The concept of PBG is also introduced to develop a high Q 2D defect metallo-dielectric resonator as shown in Figure 2.11. The reported unloaded Q for this design was 983 at 11.75 GHz on Duroid substrate [37]. In this case, a CPW feeding line (shown on top layer) is shorted directly into a slot. The slot feeds the resonator formed by removing one metallic rod.

(a) 1D-PBG bandstop filters

(b) 1D-PBG resonators

Figure 2.10. Uniplanar CPW, CPS & slot lines EBG structures [36].

Figure 2.11. Metallo-dielectric EBG defect resonator including magnitude of E field plot at resonance [37].

Finally, one might think that exploring novel 3D PBG structures could be a promising way to increase resonator Q in the future, by applying a periodicity in the third dimension. However, the challenge is in how these structures can be adapted for embedding into packaged subsystems. Especially, the characteristics of the available planar materials, the allowed number of layers, and the fabrication process still require further enhancements to be adaptable for housing 3D periodic structures [35].

2.3.4 Waveguide Resonators

A waveguide resonator can be considered as a volume enclosed by a conducting surface in which an electromagnetic wave can be excited. Perfectly conducting enclosures, or cavities, provide a means of confining energy. They are usually filled with air to get rid of any dielectric losses and obtain high Q . Therefore, they have much higher Q than any planar resonator with excellent stability over time and can be used as fixed standards. They are difficult to tune. However, they are often used in wave meters for

measuring frequency [38]. There are many types of cavity resonators as shown in Figure 2.12. But, the rectangular and cylindrical types will be specifically considered through this study.

Figure 2.12. Examples of typical cavity resonators (E lines are solid while H lines are dashed) [38].

Rectangular Waveguide Resonators are formed by a section of a rectangular waveguide of length d , as shown in Figure 2.13(a). Like other cavity resonators, this cavity can support an infinite number of modes. For higher order modes, the resonance frequency increases and Q at a given frequency also increases. Rectangular waveguide cavity resonators are known to be easily analyzed, fabricated and excited.

Circular Waveguide Resonators consist of a section of circular waveguide of radius a and length d , with two end metallic plates to provide a short circuit, as shown in Figure 2.13(b). Cylindrical waveguide cavity resonators are slightly more complicated structures to analyze, realize and mode-excite, but can offer higher Q factors (up to 30,000) due to the smaller conductor to volume ratio, in case of air filling.

Figure 2.13. Typical unloaded conventional waveguide resonators (a) Rectangular (b) Cylindrical.

The main disadvantage of cavity resonators in general is their bulky volume and hence their unsuitability for the small size requirements of many modern applications. However, dielectric filled versions of this class of resonators are very promising and can be readily implemented in packaging environments such as LTCC by replacing the vertical metallic walls by closely spaced via posts. This will be shown in the proposed designs of Chapter 3 and Chapter 4.

2.3.5 Micromachined Resonators and other related technologies:

Recent developments in microelectromechanical systems (MEMS) have demonstrated their potential in various microwave applications such as switches, phase shifters, tunable resonators and VCOs. Bulk micromachining techniques have also been used to develop high-Q resonators and filters. Cavity resonators with the Si being partly removed have shown unloaded Q's beyond 360 at 18.4 GHz [39], as shown in Figure 2.14. Some other high-Q MMW resonator structures have been designed through combining micromachining with DR and can achieve Q factors from 500 up to 2400 at 35 GHz [40].

(a) Magnified top view.

(b) photo.

Figure 2.14. Typical micromachined cavity resonator [39].

Figure 2.15. MEMS bridge Varactor.

The configuration, shown in Figure 2.15 is used as a varactor, where the electrostatic attraction draws the metal bridge down to modulate the capacitance and hence tune the resonance frequency. Because current never flows through a semiconductor material, the equivalent series resistance of a MEMS varactor can be very low. Another advantage to MEMS transducers is that they are inherently linear. So, they are very promising for tuning high-Q resonators. However, they still suffer from low maturity and high cost [41].

Finally, there are other types of resonators not discussed here (like conventionally tuned resonators with varactor diodes and YIG devices) because they are beyond the

scope of this study. But, recent developments in ferrite LTCC materials may lead to their implementation in a modern packaging environment.

It is also worth noting that superconductors are leading technologies in the field of high-Q resonators. The highest Q resonators that can be achieved from superconductors is up to 100,000 at MMW frequencies [42]. However, they suffer from cost and technological problems such as repeatable high critical temperature T_C material performance and cooling requirements.

As a summary, there is a tradeoff between the resonator Q-factor, its size, and its cost as shown in Figure 2.16. It's clear that the planar resonators can be implemented in MMIC chips [4] or in package modules [43] and hence achieving the best size and cost but they suffer from their poor Q-factors, while certain MEMS resonators can obtain higher Q with good size but have the highest cost, being the most sophisticated technology [39]. On the other hand, the conventional metallic waveguide cavities can achieve the best Q factors ever but once again with bulky size and expensive cost [44], [45]. As a compromise solution, one can choose surface mount resonators to get fair size and good Q [46] or even dielectric resonators for better Q [47] but both have considerable assembly cost. Therefore, this thesis will try to find an alternative low cost, miniaturized, high-Q resonators, as will follow in Chapter 3 and Chapter 4.

Figure 2.16. Performance comparison between the most common MMW resonators and the proposed structures.

2.4 Conclusion

In this chapter, the concept of resonance and the definition of the performance parameters are introduced. Also presented in this chapter is a survey of the available resonators at microwave and millimeter-wave frequencies, along with their technology suitability, advantages and disadvantages. Finally, their methods of analysis and design as well as measurements are discussed.

Planar resonators are clearly the most convenient to fabricate, but they have the lowest Q factors (less than 300). Waveguide cavity resonators allow higher Q values (up to 30,000) but with bulky dimensions. Dielectric resonators (DR) can also offer as high Q as (20,000) but they are not amenable to mass fabrication due to their frequency-sensitivity. The highest Q resonators can be achieved from superconductors (up to

100,000). However, they suffer from cost and technological problems such as repeatable high critical temperature T_C material performance and cooling requirements. This suggests that the LTCC housing of an entire subsystem could be a promising route to explore novel high Q structures and embed those resonators with the MMIC chips, as will be seen in the following chapters.

CHAPTER 3 Proposed High-Q LTCC Resonators For Millimeter-Wave Applications

3.1 Introduction

During the last few years, LTCC has emerged as a high performance packaging technology capable of embedding a microwave circuit's passive elements and interconnects. In pursuit of the first and second thesis objectives, this chapter investigates LTCC-based resonators that provide high quality factors at frequencies extending into the MMW band. The chapter begins with a brief overview of LTCC technology; its advantages, challenges as well as its process flow diagram. Then, some novel LTCC resonator configurations are proposed, which are divided into rectangular and cylindrical geometries. The resulting multilayered structures are called Laminated Wave Guide (LWG) resonators or Via Wave Guide (VWG) resonators.

For the rectangular resonators, various feeding mechanisms (probe, loop, coplanar line) and filling materials (air, dielectric) are experimentally investigated. A design strategy for obtaining high Q factors using a number of loss-reduction and feeding techniques is proposed. Finally, a few optimum designs are carried out, yielding large unloaded Q values. Due to their unique properties, cylindrical resonators will be studied in detail in Chapter 4.

3.2 An Overview of LTCC Packaging Technology

Basically, Low Temperature Co-fired Ceramics provide multi-level (multi-layer) packages. These packages can reserve levels for DC routing, levels for RF and microwave signals, and several layers for buried lumped elements, antennas, sensors, etc. An example of an LTCC-based microwave transceiver circuit is shown in Figure .3.1. LTCCs allow multiple active circuits to be mounted on the same package substrate along with all their attendant off-chip passive or even radiating components. Therefore, LTCCs can not only create complete systems in a package, but also eliminate the need to go off-package for any components and hence improve circuit performance [48], [49], [50].

Figure .3.1. An example of a multi-layered LTCC module showing some of its integration capabilities [51].

The LTCC manufacturing process is summarized in the flow diagram of Figure 3.2. In this automated process the green dielectric tape is first cut into blanks for all layers. The vias are punched in these blanks and filled with ink (silver paste, for example). This is followed by the screen-printing of the conductors on all layers. The printed blanks are collated with orientation guide and laminated under certain pressure. In the last step, the laminated blanks are co-fired at a temperature of about 850° C. Discrete chips may be housed on lower layers by forming cavities. For such packages, small

metallic boxes are inserted in all air cavities before co-firing to protect the ceramic walls from being corrugated or collapsed.

Figure 3.2. LTCC process flow diagram [52].

3.2.1 LTCC Advantages and Challenges

LTCC has many advantages, which can be summarized as follows: Multilayer (3D) circuitry is allowed, whereby the vertical dimension may be exploited, leading to very compact circuits and modules. A high power handling capability exists due to large dielectric breakdown voltages. Its high layer count (>50 layers), allows modules of up to 3.5 mm in thickness to be realized with a large surface area (up to 100 x 100 mm). LTCCs are environmentally insensitive to humidity, heat, cold, and shock. They provide excellent high frequency performance due to low loss conductors and dielectric materials. In addition, multi-level cavities with different sizes may be inserted, allowing hybrid

technology integration. Finally, passive components may be embedded (buried R, L, and C) within the module, allowing substantial size reduction of the active chips.

On the other hand LTCC has some challenges. LTCC can be separated into two sub-categories: LTCC, and LTCC-M (LTCC on Metal). One disadvantage of LTCC is that it shrinks in the x, y, and z directions upon firing. Because of the “xy” shrinkage, wide solid ground planes are not possible. On the other hand, LTCC-M can offer better performance because the ceramic layer attach to the metal and prevent the ceramic from shrinking in the x and y directions and allows solid ground planes to be fabricated easily.

Shrinkage in ‘z’ can lead to some defects, such as 'posting', where a via remains the same height but the layer through which it passes shrinks, causing the pads that meet the vias on the layers above and below to be pushed out of their optimum position. This is due to the fact that the vias are filled (referred to as solid vias), and the conductive ink that fills the via does not shrink along with the ceramic. The disadvantage of LTCC-M is its large “z” shrinkage with high uncertainty. LTCC DuPont 943 has “z” shrinkage of 12.3 +/- 0.3%, while the DiPak LTCC-M process has “z” shrinkage of 30+/- 10% [53]. Therefore, LTCC-M will not be used to produce the prototypes required for this thesis, since the z direction is an important parameter in the proposed designs.

Generally speaking, LTCC packages are poor conductors of heat compared to laminated multi chip modules (MCM_L). However, thermal conductivity can be improved by using thermal vias to carry heat from hot microwave semiconductor devices to a metal core or a heat sink mounted on top of the package. Also, the coefficient of thermal expansion (CTE) of the package material becomes critical. The CTE of a

material is defined as the fractional increase in length per unit rise in temperature. The CTE of most commercial brands of LTCC and LTCC-M is set to match closely with GaAs so that the active device remains attached to the package [54].

3.2.2 Further Promising Issues in LTCC

Recently, dielectric-filled rectangular waveguides have been built in LTCC-M, with some appropriate modifications such as synthesizing the narrow walls of the guide with rows of vias [55]. This type of guide will be referred as via waveguide (VWG). Metal laminations can also be used on the layers to bridge the vias, forming a meshed wall. This type of guide will be referred to as laminated waveguide (LWG). The LWG has good isolation from a neighboring LWG, and it has been confirmed that fundamental structures, such as bends, branches power dividers can be realized with good performance [56], [55]. These basic waveguide structures are adapted through the thesis such that new resonators may be designed, analyzed and fabricated using the accessible LTCC foundry facilities.

A high dielectric constant is known to lower the dimensions of the LWG. This could open the way to explore normally size-consuming configurations like the promising EBG periodic structures, which were discussed in Chapter 2. However, being a more recent product offering, high dielectric LTCC materials are not openly disclosed by their suppliers.

Nowadays, ferrite LTCC materials, both pastes and tapes, are known to exist and have been used in very recent applications such as performance improvement in multi-layered inductors and mm-wave phase shifters [57]. This should encourage the mm-wave

designers to investigate the characteristics of the ferrite LTCC materials and employ them to tune the mm-wave oscillators. If successful, it may be very promising to utilize the LTCC for current controlled oscillators and revive all the advantages of the traditional YIG-tuned oscillators.

One other issue is the packaging of high-Q micromachined varactors (MEMS varactors). If the MEMS varactors are mature enough and can be embedded in the same LTCC module with the MMIC oscillator, then an ultra low phase noise VCO could be implemented.

3.3 Configurations of LTCC Waveguide Resonators

Conventional rectangular and cylindrical waveguide resonators were discussed in Chapter 2 and were shown to exhibit larger quality factors than planar resonators. Given the 3D circuit fabrication capabilities of LTCC, dielectric filled versions of these two classes of volume resonators are proposed for implementation in LTCC. This will be done by replacing the vertical metallic walls by closely spaced via posts, as shown in Figure 3.3. Such a structure has been employed in the past, but only for a low frequency resonator design, which appears to have been abandoned due to its poor Q of approximately 150 at 5 GHz [16].

The idea of using via posts to form a resonator was also used in [58] to implement a Duroid substrate-integrated waveguide (SIW) having a Q above 500 at 21 GHz. In addition, via posts have been used to form photonic band gap (PBG) resonators in Duroid substrates with Q above 700 at 9 GHz [61]. Although these values of Q are impressive,

Duroid is not a packaging technology. Furthermore, it will be shown here that LTCC-based VWG resonators can achieve even higher quality factors.

All proposed designs are implemented using a standard LTCC process. The Ferro A6-S LTCC tape system is chosen, having $\epsilon_r=6.2$, and $\tan\delta=0.0013$, which is one of the lowest loss LTCC materials. Since the LTCC material parameters may depend on the used foundry, they can not be obtained directly from the data sheet, instead they were chosen based on the measurements shown in Appendix A. Minimum sized via diameters and spacing are employed. The module thickness is 0.8 mm, which corresponds to 8 LTCC layers of 100 μm each. All proposed designs are modeled, simulated, and verified with measurements. For each case study, many tradeoffs and guidelines are highlighted for the designer's benefit.

3.3.1 Design Considerations

The design of LTCC waveguide resonators closely follows that of conventional dielectric filled rectangular or cylindrical waveguide resonators. For a ceramic filled resonator with via walls, the resonant frequency can be obtained based on the effective rectangular waveguide cavity dimensions proposed by Cassivi et al [58], [59] which is valid when the via pitch p is kept sufficiently small. For p values greater than 5 times via diameter, Cassivi's method cannot be applied. Thus, a theoretical approach has been derived to predict the effective dimensions of such laminated waveguide cavities [60]. Once the effective dimensions are obtained, the conventional rectangular waveguide design equations can be used to obtain the resonant frequencies. However, 3D EM

simulations are more suitable to explore complex laminated structures with different excitation techniques.

The recent advent of general-purpose three-dimensional (3D) electromagnetic commercial codes, has stimulated interest in developing new microwave circuit and antenna structures. In general, there are many different methods of full wave simulation, which can be employed such as finite-difference-time-domain (FDTD) techniques, finite element methods (FEM), as well as moment methods. Each of these techniques has advantages over the others. For example, FDTD techniques can provide the transient response of a device, and with the FFT, frequency domain data can be obtained. However, the simulation is extremely time consuming, especially in the case of resonant structures. Frequency domain techniques, such as FEM and moment methods offer faster run times, however, they only provide data at a single frequency and therefore must be repeated over many frequencies. Additionally FEM requires the discretization of the entire frequency domain to absorb large computer memory.

On the other hand, the time-domain Finite Integration Technique (FIT) offers the advantage of requiring much less memory and computing time [62]. It provides a universal spatial discretization scheme, applicable to various electromagnetic problems including 3D resonators. The 3D structure is excited with a broadband wave and solved in the time-domain. Then an FFT yields the S-parameters for the whole frequency range with just one calculation run. This not only saves time, but also avoids the risk of missing sharp resonance peaks (unlike with frequency calculations). Therefore, this thesis makes use of a commercially available program called CST Microwave Studio, which is based on FIT, to model the proposed 3D resonant structures. The mesh strategy of the 3D EM

models is based on the Perfect Boundary Approximation (PBA) technique [63]. This technique offers the advantage of allowing an accurate representation of curved structures.

3.3.2 Basic Resonator Designs

As a starting point for future reference, two basic LTCC resonators - a rectangular geometry and a cylindrical one - are designed for 23 GHz operation. A 3D EM model is built for each and the eigen modes are analyzed. The designs with their analytical results are summarized in Table 3.1. f_o is the resonance frequency of the first resonance mode for each and Q_o is the corresponding unloaded quality factor.

Table 3.1
Rectangular versus cylindrical VWG resonators in LTCC

Parameter	Rectangular	Cylindrical
Hight [μm]	700	700
Via Pitch p, [μm]	400	400
Via diameter phi, [μm]	150	150
Resonator Size, (axb) [mm^2] or Diameter (2r) [mm] Total used Area [mm^2]	3.7x5.7 21.09	3.8 11.35
1 st Resonance Mode	TE ₁₀₁	TM ₀₁₀
f_o [GHz]	23	23
Q_o	320	275

As can be seen, the dielectric-filled VWG cylindrical resonator has slightly lower Q than the rectangular one. However, the main advantage of cylindrical resonators is their small volume compared with the rectangular ones. Occupying approximately half the area, they are much less expensive to produce. A more detailed treatment of the cylindrical resonators will be given Chapter 4. Also note that a fixed size resonator can be operated at higher order modes to obtain higher resonant frequencies. For simplicity, the

results of Table 3.1 are based on an eigen mode analysis which does not take into account the resonator's excitation method.

The Q values obtained are still quite low for both configurations of Figure 3.3, due to high dielectric and radiation losses. Some Q enhancement techniques can be employed like inserting an air cavity and/or applying staggered double via walls to overcome those drawbacks. Also, practical feeding mechanisms need to be examined. These points will be addressed in the following sections.

(a) Rectangular configuration

(b) Cylindrical configuration

Figure 3.3. Basic VWG structures of LTCC resonators.

3.4 Novel Excitation Techniques for LTCC resonators

For multi-layered rectangular or cylindrical resonators implemented in LTCC technology, some possible excitation techniques are shown in Figure 3.4. Note that all the designs illustrated employ an air cavity to improve Q , as will be discussed later in detail. The effects of feeding mechanism on resonant frequency and quality factor are as follows:

3.4.1 Probe Excitation

An LTCC resonator may be excited by a small waveguide probe that consists of a via descending into the guide through a small circular aperture in the upper broad resonator cover as illustrated in Figure 3.4(a). This probe via may go through only one layer of the LTCC tape so as to be much smaller than the operating wavelength to achieve loose coupling [64]. Also, for the same reason, it is recommended that the probe be away from the maximum of the electric field of the desired mode of resonance. Therefore the loaded quality factor Q_L will approach the value of unloaded factor Q_o .

3.4.2 Loop Excitation

Loop coupling may be employed in LTCC as shown in Figure 3.4(b). The size and location of the loop with respect to the H field for each mode determines the degree of coupling. For both loop and probe excitations the resonance frequency and/or the unloaded quality factor Q_o may be affected due to the loop/probe position and its coupling depth. Four VWG rectangular resonators have been designed and fabricated with the same parameters but different loop positions and lengths in order to investigate how much that dependence will be. Although not mentioned here, an optimum loop feed

design consists of a 1-layer depth via with a small strip length at the narrow wall side “a” gives better Q and less frequency drift than other longer and deeper loop designs at this side or at other sides [65].

(a) Probe Excitation

(b) Loop Excitation

(c) A novel excitation using center strip of a CPW

Figure 3.4. Possible excitation techniques for LTCC Resonators.

3.4.3 Excitation Using Center Strip of a CPW

While the probe and loop feeds are simply adaptation of conventional waveguide design to LTCC, Figure 3.4(c) shows a novel excitation using the center strip of a CPW. The strip is straight and longitudinal but closer to one wall. The entire cavity is excited by the TEM wave of the strip. The advantage of this method is that flip-chip mounted MMICs in neighboring cavities within the package may be connected easily and with minimal discontinuities or vertical transitions. It is worth mentioning that vertical transitions in LTCC packages may cause significant loss of MMW signal. So, avoiding them is a significant advantage for the overall package design. Also, tuning of resonance modes using the strip width and length is much easier than with loops or probes, because they are not constrained by discrete values of via-diameters and layer thicknesses.

A potential disadvantage of CPW strip excitation is the excitation of a radial-like mode inside the resonator, which can resonate close to the fundamental TE_{101} mode. This mode is shown in Figure 3.4(c) and can be tuned out with proper care by the designer.

Figure 3.5. A photograph of a typical CPW-fed rectangular resonator [65].

3.5 Proposed Q Enhancement Techniques

Several configurations of LTCC resonators have been designed, fabricated and measured, and the results are summarized in Table 3.2 and Table 3.3. Analysis of these designs leads to the following proposed techniques to maximize the quality factor of LTCC resonators:

1) Increasing the height, b , of the resonator will increase the resonator volume and hence increase its Q . It is therefore recommended to include as many layers as possible in the resonator. The number of layers was chosen here to be only 8 as a compromise between low cost and high Q . On the other hand, altering ' b ' does not affect the resonance frequency. For a better understanding, assume a rectangular waveguide (RWG) resonator in the TE_{101} mode. Equation 2.9 can be evaluated using the effective resonator dimensions. Then the result of conductor loss Q_c , will be as follows [66]:

$$Q_c = \left[\frac{k^3 abd\eta}{4\pi^2 R_s} \right] \left[\frac{1}{[l^2 ab/d^2 + bd/a^2 + l^2 a/2d + d/2a]} \right] = F_1 F_2 \quad (3.1)$$

Where: F_1 is the first term in the first brackets and F_2 is the term in the second brackets.

It is clear that for ' b ' values much smaller than a and d (as is the case with LTCC), the effect on F_2 is not noticeable. On the other hand F_1 has b directly multiplied by a and d . So, it will increase F_1 much greater than its degradation effect on F_2 . Hence, the conductor component of quality factor will increase with increasing b .

2) Double rows of staggered vias, rather than a single wall of vias, can be used as sidewalls in order to minimize the resonator radiation loss. Such a scheme is employed in Figure 3.6. This technique is particularly useful for high-Q air-filled structures where Q_r is no longer negligible. So, adding another row of vias staggered with the first one will increase the Q_r by containing more fields inside the resonator and hence minimizing the radiation dramatically. In order to prove this technique experimentally, four resonators were designed, fabricated and measured; one probe-fed, one CPW-center-line-fed, and two having different loop-fed designs. The measured resonant frequencies and associated Q values are shown in Table 3.2. It can be noticed that at least 50% enhancement is achieved for Q_o in loop-fed and probe fed cases. But for the CPW-fed resonator, the increase is limited due to the absence of a double via wall at the feeding side.

On the other hand, attention must be paid to the resonant frequency since it will change slightly with the addition of the second row. In this case, the effective dimensions of a double via sidewall resonator are smaller than those of a single sidewall structure due to the fields being more tightly contained. Therefore, the resonant frequency will increase, as verified by the measured data shown in Table 3.2.

Table 3.2
Comparison between resonators with single sidewalls of vias and those of double staggered via walls

	Probe-Fed P1S&D		Loop-Fed, 1st L1S&D		Loop-Fed, 2nd L2S&D		CPW-Fed CPW2,CRC,N&M	
	f_o [GHz]	Q_o	f_o [GHz]	Q_o	f_o [GHz]	Q_o	f_o [GHz]	Q_o
Single wall	23	196	23.2	205	21.1	213	35.8	230
Double staggered walls	23.3	300	23.3	315	21.27	320	36.2	270
% increase	1.3	53.1	0.4%	53.6	0.8	50.2	1.1	17.4

3) Using a ceramic material with the lowest possible loss tangent will obviously increase Q_d and hence the overall Q_o . The limiting case of this is to put an air cavity inside the LTCC resonator, which will reduce the dielectric losses dramatically. The resonator must then be redesigned in order to obtain the same resonant frequency due to the change in the dielectric constant. As shown in Table 3.3, a typical LTCC resonator having $Q_o = 330$ at 38 GHz can exhibit a Q_o of 540 when it is redesigned and resimulated with an appropriate air filling. It is clear from the Table that the enhancement in Q_o is greater than 50% except for one case, proving the effectiveness of this technique.

Table 3.3

Measured f_o and Q_o for dielectric filled LTCC resonators versus those of resonators with Air cavities

		Without Air Cavity			With Air Cavity			Enhancement
		Size [mmxmm]	f_o [GHz]	Q_o	Size [mmxmm]	f_o [GHz]	Q_o	ΔQ_o [%]
Loop-Excited Rectangular Resonator	1 st Mode	L2d 3x5 mm	21.27	335	L1C 9x11, with 8x10 cavity	19.4	630	88
	2 nd Mode		28.9	370		32	632	71
	3 rd Mode		37.9	330		38	540	64
Probe-Excited Rectangular Resonator	1 st Mode	P1d 3.3x5.25 mm	23.3	300	P2CII 7.5x9, with 6.5x8 cavity	23	400	33.3
	2 nd Mode		31.5	270		33.8	500	85

4) Finally, a few other practical issues affect the quality factor. For instance, each LTCC layer has a catch pad for every via post position. The catch pads of these vias may be connected with a strip line to enhance the conductive area of the sidewalls by about 8%. As a result, the Q_r will increase while Q_d will not be affected noticeably. So, the overall Q_o will be higher. Also, from the point of view of achieving a high loaded quality factor Q_L , the coupling between the resonator and its feed line should be sufficiently weak by

reducing the feeder size and by shifting its position away from the point at which the field of interest is maximum. In this case, Q_L may approach Q_o .

In summary, the above points show there are many factors which should be taken into consideration to obtain a high-Q LTCC resonator. Along with the excitation techniques of the previous section, a number of design guidelines have now been presented. Although mentioned one-by-one in this section, the techniques will be used in combination to produce some novel designs in the following section.

3.6 Novel High-Q LTCC Resonators

Based on the preceding discussion, three optimum designs have been carried out in LTCC. These are rectangular resonators with different feed types and containing air cavities. The material parameters were given in Section 3.3.

The first optimized design is shown in Figure 3.6. A microstrip-connected probe feed is used to excite the resonator from the backside metal in order to reserve as many layers as possible for the resonator height b . Double staggered via rows are used as sidewalls to increase Q . For each layer, the catch pads of vias are connected with a strip line, as shown (see also Figure 3.4). The air cavity is created inside the resonator with four layers depth and with the maximum allowed size that respects the LTCC fabrication rules. In this case, the optimized resonator dimensions are: $a = 8\text{mm}$, $b = 0.7\text{mm}$, $d = 9.9\text{mm}$, via diameter $\phi = 150\mu\text{m}$, and via pitch $p = 0.4\text{mm}$, with an air cavity of $7\text{mm} \times 9\text{mm}$. This novel LTCC resonator shown in Figure 3.6 and its experimental results are presented in Figure 3.7. The predicted and measured resonant frequencies as well as the Q values are reported in Table 3.4. The unloaded Q is extracted from the measured

Scattering parameters using the method described in Appendix C (as will be the case for all other resonators). While agreement with predicted values is excellent for the TE_{102} mode, noisy measurement results prevented certain resonance modes from being observed (TE_{201} , TE_{202}) or did not allow their Q values to be extracted with high accuracy (TE_{101} and TE_{103}). It is clear that very high-Q values (>1000) have been achieved. In fact, this LTCC resonator is believed to be the first design showing such a high Q in a standard LTCC process for mm-wave applications [68].

Figure 3.6. Layout of an LTCC Resonator with double staggered via walls and Air Cavity.

Figure 3.7. Measured S11 in dB for the novel probe-fed rectangular resonator of Figure 3.6.

The third optimized design is also a loop fed resonator but the loop is positioned differently to excite the TE_{102} and TE_{202} modes, which did not appear in the second optimized design. It shows the highest Q obtained for a loop-fed rectangular LTCC resonator, which is 700.

Table 3.4
Measured Versus Simulated Q Values for LTCC Resonators with Air Cavity

Mode	Predicted (3D EM model)		Probe-fed Rect. Resonator Measured				Loop-fed Resonator1 Measured				Loop-fed Resonator2 Measured			
	f_o [GHz]	Q_o	f_o [GHz]	Q_o	Δf_o [%]	ΔQ_o [%]	f_o [GHz]	Q_o	Δf_o [%]	ΔQ_o [%]	f_o [GHz]	Q_o	Δf_o [%]	ΔQ_o [%]
TE_{101}	20.276	1022	19.81	1206*	-2.3	18	19.44	630	-4	-38.4	-----	----	-----	----
TE_{102}	29.053	1054	28.672	1102	-1.3	4.5	-----	-----	-----	-----	30.27	700	4.1	-33
TE_{201}	32.586	1013	-----	-----	----	---	32.1	632	-1.5	-37.6	-----	----	-----	----
TE_{202}	37.653	982	-----	-----	----	---	-----	-----	-----	-----	36.4	550	-3.2	-44
TE_{103}	38.26	886	38.674	844*	1.1	-5	38.1	540	-0.4	-39	-----	----	-----	----

* Uncertain result due to noisy measurement.

A photo of the fabricated LTCC High Q resonator showing a footprint for future placement of a MMIC flip chip, is given in Figure 3.8.

To summarize, the design techniques described in this chapter have proven to be effective in achieving the highest-ever reported Q factors for LTCC-based MMW resonators. The probe-fed design yields the highest Q. Excitation of specific modes can be controlled with appropriate choice of feed type and location. The loop excitation will never produce as high Q factors as the probe-fed designs due to incomplete closure of the loop in fabrication.

However, improved loop implementations are likely to lead to higher Q. Finally, CPW excitation offers significant advantage in terms of flip-chip mounting for neighboring circuits and reserves more layers for building up the resonator thickness, as required for large Q.

Figure 3.8. Photo of the fabricated LTCC High Q resonator showing a footprint for the MMIC flip chip.

3.7 Conclusion

In this chapter, high-Q LTCC resonators have been investigated for mm-wave applications. Our modeling and simulation methodology has been experimentally validated through many different LTCC resonator designs at mm-wave frequencies. A design strategy for obtaining high Q-factors has been proposed, which includes the insertion of air cavities and novel excitation structures along with design guidelines.

Based on these techniques, three novel LTCC high-Q resonators were designed, optimized and fabricated. Measured unloaded Q of 1100, 700 and 550 were obtained at Ka band, for probe-fed rectangular and different loop-fed rectangular configurations, respectively. These values are the highest ever reported for a standard LTCC process and fulfill the first two thesis objectives. It is believed that the proposed resonators will be very useful in the design of low-cost mass-producible MMW filters and oscillators. Demonstration of the novel high-Q resonators in enhanced performance MMW oscillators is provided in chapter 5.

CHAPTER 4 Study Of Cylindrical Multilayered Ceramic Resonators With Rectangular Air Cavity

4.1 Introduction

Chapter 3 focused on LTCC-based resonators of rectangular geometries capable of providing high quality factors. It was seen that, among the other techniques, the removal of lossy dielectric material within the resonator leads to substantial increases in Q . Due to the current limitations of LTCC fabrication technology, such air cavities can only be made rectangular in shape. Thus, the realization of a very high Q cylindrical resonator in LTCC requires that it contain a rectangular air cavity, but one that can only partially fills its volume. What results is a completely new structure which has not been encountered before in open literature. The cylindrical resonator is nonetheless very important since it allows an almost 50% size reduction compared to rectangular LTCC resonators. This chapter is devoted to the study of these novel cylindrical resonators, in fulfillment of the third major thesis objective. Section 4.2 outlines the basic problem and its modeling approach. The main effects on resonator performance of various filling factors and excitation methods are then presented in Sections 4.3 and 4.4, respectively. Finally, an optimum high Q cylindrical resonator implementation is described in Section 4.5.

4.2 Design and Modeling of Via Wave Guide LTCC Resonators

The fundamental TM_{010} and the TM_{nlp} higher order modes are shown in Figure 4.1 for a cross-sectional view of a conventional cylindrical resonator with radius ' a ' and thickness ' d '. The electric field lines are solid while the magnetic ones are dashed. The white spots refer to maximum E-field density areas.

Figure 4.1. A conventional cylindrical cavity resonator field distributions for the lowest 6 TM modes [108].

The analytical expressions for the field components, modal resonant frequencies and Qs have been derived in several textbooks [38] and will be simply stated here for the first mode TM₀₁₀.

$$E_z = E_0 J_0\left(\frac{2.405r}{a}\right) \quad (4.1)$$

$$H_\phi = \frac{jE_0}{\eta} J_1\left(\frac{2.405r}{a}\right) \quad (4.2)$$

$$f_o = \frac{2.405}{2\pi a \sqrt{\mu\epsilon}} \quad (4.3)$$

$$Q_c = \frac{\eta}{R_s} \frac{t_{01}}{2(1 + a/d)} \quad (4.4)$$

$$Q_d = \frac{\omega W}{P_d} = \frac{\epsilon'}{\epsilon''} = \frac{1}{\tan \delta} \quad (4.5)$$

where: $\eta = \sqrt{\mu/\epsilon}$,

$J_n(.)$ is the n th-order Bessel functions of the first kind,

$t_{01} = 2.405$ is the first nonzero root of $J_0(.) = 0$,

R_s is the surface resistance of the metallic cavity conductor,

and $\tan \delta$ is the loss tangent of the dielectric filling.

For the proposed LTCC cylindrical resonator structures, a 3D illustration is given in Figure 4.2 with the following parameters:

d_1 and d_2 are the centre to centre diameters of the inner and outer via walls, respectively.

p is the via pitch; the spacing between 2 successive vias from centre to centre.

Figure 4.2. A 3D illustration showing (a) the vertical parameters and (b) a 3D model for the novel LTCC resonator structure with rectangular air cavity in the middle and double rows of silver via posts act as a cylindrical wall.

A is the side length of the square air cavity.

the length of the cylinder L is given by:

$$L = l_{total} * t \quad (4.6)$$

where t is the layer thickness and l_{total} is the total number of layers:

$$l_{total} = l_{diel} + l_{air} \quad (4.7)$$

where l_{diel} is the number of dielectric layers, while l_{air} is the number of layers used to form the air cavity. Another important design parameter is the air cavity depth ratio which is defined as

$$d = \frac{L_{air}}{L_{total}} = \frac{l_{air}}{l_{total}} \quad (4.8)$$

Finally, the relative dielectric constant ϵ_r is defined to be one inside the air cavity and greater than one elsewhere in the LTCC dielectric region.

To rigorously treat this problem analytically would require enforcement of a double set of field boundary conditions at the air/dielectric and dielectric/metallic interfaces with appropriate conversion between cylindrical and rectangular coordinates [70]. Such an undertaking is beyond the scope of this Thesis, and a simple approach will be followed here which yields more practical results.

Based on numerical EM CAD results, the material parameters $\tan\delta$ and ϵ_r of the dielectric, and the geometric parameters A , d , a and f will be mapped to the expressions describing the conventional resonator of Figure 4.1. Thus Equations (4.3) to (4.5) may be subsequently employed to predict the performance of arbitrary LTCC cylindrical resonator with rectangular air cavities. It is critically important to understand the trade-offs between size, air/dielectric filling ratios and Q, as faced by the designer.

Implicit in the above approach is the assumption that the presence of the air cavity does not completely alter the original field configurations inside the resonator; an assumption that will be presently validated. This allows the concept of an effective dielectric constant $\epsilon_{eff} = f(A, d, L, a, \epsilon_r)$ and an effective radius $a_{eff} = f(d_1, d_2, p)$ to be defined and employed in lieu of ϵ and a in Equations (4.3) to (4.5).

The following analyses are based on LTCC VWG cylindrical resonators implemented using the Ferro K6S LTCC tape system, having $\epsilon_r=6.2$, and $\tan\delta=0.0013$, which is one of the lowest loss LTCC materials. The module thickness is 0.8 mm, corresponding to 8 LTCC layers each of $t=100 \mu m$ cofired thickness. CST microwave Studio is again employed, as in Chapter 3. The 3D structure is simplified to that shown in Figure 4.2 (b) for the studies throughout the rest of this chapter. It consists of 8 layers, seven of which are used to build the resonator height, while one layer is reserved for the

probe or loop excitation. But, as mentioned before, the inner conductor of a CPW feed would not use this eighth layer. This model makes use of the advantage of building the resonator wall with double staggered via rows. The structure is modeled as an eigen value problem to study the natural resonance frequencies of the structure and their associated Q values, without introducing a physical port excitation method.

Figure 4.3 and Figure 4.4 show the calculated E-field and H-field magnitudes for the fundamental TM_{010} mode of a family of resonators with increasing air cavity side length A . All field magnitudes are normalized to the maximum field values of the case of air-filled equivalent cylindrical cavity resonator as follows:

$$|\overline{E}(x)| = \frac{E(x)}{E_{o\max}(x)} \quad (4.9)$$

$$|\overline{H}(x)| = \frac{H(x)}{H_{o\max}(x)} \quad (4.10)$$

It is clear that the air/dielectric interface creates a discontinuity in the field distribution across the structure, but its effect is relatively minor. For instance, no new field maximum is introduced by the heterogeneous filling of the resonator, and the field peak locations exhibit only a slight displacement in the case of the H-field. Based on these observations, it is possible to pursue with effective ϵ_{eff} assumption [67]. A more detailed discussion of the VWG resonator's field behavior is given in Section 4.5.

Before going into detailed parametric analysis for the proposed resonator structure, two important parameters must be defined to relate the proposed novel structure to the conventional dielectric-filled solid-wall cylindrical resonator.

Figure 4.3. A comparison between proposed VWG and conventional dielectric-filled cylindrical resonator.

Figure 4.4. Magnetic field distribution along the x-axis for different air filling volumes.

a_{eff} : is the effective radius of the proposed dielectric-filled cylindrical VWG resonator, which replaces the radius a of an equivalent conventional dielectric-filled solid-wall cylindrical resonator having the same first resonance frequency f_{o1} . Since $2a_{eff}$ may be somewhere in the range of d_1 to d_2 (see Figure 4.2) it is necessary to determine this quantity exactly.

ϵ_{eff} : is the effective relative dielectric constant of the proposed cylindrical VWG resonator with partial air filling. ϵ_{eff} therefore replaces ϵ_r of an equivalent conventional dielectric-filled solid-wall cylindrical resonator having the same first resonance frequency.

First, the effective radius is determined, by analyzing a VWG resonator and comparing it with a few dielectric-filled solid-wall cylindrical resonators with different diameters in the range d_1 to d_2 , based on Equation (4.3). The results are shown in Figure 4.5.

Figure 4.5. A comparison between a dielectric-filled VWG resonator with double rows of vias and few solid-wall dielectric-filled resonators.

It is immediately seen that the via wall resonator curve is superimposed on the curve corresponding to a solid wall resonator with $2a=d_1$. Thus $2a_{eff}$ is taken to be the centre to centre spacing of the innermost row of vias forming the VWG resonator walls. All subsequent analyses are based on VWG structures having this same a_{eff} value.

Second, the ϵ_{eff} is determined. The resonant frequency of a dielectric filled solid-wall resonator is calculated using Equation (4.3) as a function of ϵ_r , and is plotted in Figure 4.6. Next, the resonance frequencies of constant diameter VWG cylindrical resonators are found for various sizes of rectangular air cavities within the structure. The results are then mapped to the curve in Figure 4.6 so as to extract an equivalent ϵ_r for the VWG resonator. Figure 4.7 plots ϵ_{eff} for different d and A values, where d is the cavity depth ratio defined by Equation (4.8) and A is the cavity side length as illustrated in Figure 4.2.

As expected, the larger the air-filling ratio in the resonator (large A and d values), the lower is ϵ_{eff} . It is interesting to note that for a typical LTCC ceramic material system having $\epsilon_r = 6$, the effective dielectric constant can vary in the range of about 1.5 to 5.

The family of curves shown in Figure 4.7 may be perfectly described by a relatively simple mathematical expression relating the air filling parameters and the dielectric constant of the equivalent conventional cylinder. Curve fitting yields the effective relative dielectric constant as a function of A and d within the stated limits of each parameter to be:

$$\epsilon_{eff} = \frac{1}{0.128 + 0.09A^2} - d - 0.7786 \quad (4.11)$$

Figure 4.6. The first resonance frequency of the equivalent solid-wall cylindrical resonator versus its dielectric constant variable.

Figure 4.7. The effective relative dielectric constant of the VWG resonator of Figure 4.2 versus its air filling parameters (A and d); dots are taken from simulation while lines are the fitting curves of Equation (4.11).

Comparison of the EM simulated results (plotted as dots in Figure 4.7) with Equation (4.11) (solid curves) shows excellent agreement.

Perhaps of greatest significance to the designer is that the now established relationship for a_{eff} and ϵ_{eff} allow rapid determination of resonance frequency for arbitrary VWG cylindrical resonators, simply by using Equation (4.11). This represents a tremendous saving in computational effort compared to full 3D EM simulation.

Additional design parameters of this novel structure will be studied along with its resonance behavior in the next section.

4.3 Parametric analysis

In order to realize an optimized design for a prescribed Q, an investigation of the parameters affecting the proposed resonator performance will be conducted. This parametric analysis will study the non-homogeneous problem of cylindrical VWG LTCC resonators containing a rectangular air cavity, as shown in Figure 4.2. As before, all structures under consideration have diameter $d_1 = 7$ mm.

4.3.1 Resonance Frequency

Changing the air filling ratio within the resonator was seen to alter ϵ_{eff} . This in turn affects the resonant frequencies and the corresponding Q factors of the different modes. Figure 4.8 plots the frequencies of the lowest 2 VWG resonances. It shows that the frequency increases monotonically as the air filling increases, in a mildly exponential fashion for the first mode and in a higher exponential fashion for the second mode.

(a) first resonance.

(b) second resonance.

Figure 4.8. Frequency variation due to changing the cavity size inside the Cylindrical VWG resonator; for the lowest two resonance modes; f_{01} and f_{02} .

One may easily perform curve fitting to obtain the following relationships of the first resonance mode as a function of air filling parameters A and d :

$$f_{o1} = A^{1.54 + \frac{0.4}{7}} + 11.4 \quad \text{for } d=4/7 \quad (4.12)$$

$$f_{o1} = A^{1.54 + \frac{0.8}{7}} + 11.4 \quad \text{for } d=5/7 \quad (4.13)$$

$$f_{o1} = A^{1.54 + \frac{1.2}{7}} + 11.4 \quad \text{for } d=6/7 \quad (4.14)$$

Where, A is in mm and f_{o1} is in GHz. Combining the three equations gives:

$$f_{o1} = A^{1.54 + 0.4 \frac{(4 - l_{diel})}{l_{total}}} + 11.4 \quad \text{for } l_{diel} = 1, 2 \text{ or } 3 \quad (4.15)$$

Figure 4.9 shows excellent agreement has again been obtained between the analytical expression (solid curves) and the EM computed results (dots).

Figure 4.9. Simulated first-mode resonance frequencies (dots) and fitting curves (lines) versus air volume parameters (A and d).

Equation (4.15) describes the relationship between the first-mode resonance frequency and the LTCC design parameters (A , l_{diel} , l_{air}) for the special case of $\epsilon_r=6.2$. A more extensive study could be easily incorporate ϵ_r as a variable. Also, limiting l_{diel} to be either 1,2 or 3 may seem like another special case, but these numbers are selected on the basis of the current practical foundry guidelines. The trade-off is between putting l_{diel} equal to zero, which will give the best quality factor, and putting a few dielectric layers (like 1,2 or 3) to prevent warping.

4.3.2 Loss and quality factor

Again, the number of layers and the side-length of the square air cavity are changed to investigate the quality factor of the first mode, yielding the curves of Figure 4.10. Based on its clear exponential behavior, an exact curve fit is obtained with the relationship:

$$Q_{01} = A^{3.8+0.1d} + 380 \quad (4.16)$$

Which is valid for $\epsilon_r=6.2$ over the limits of A and d indicated. Excellent agreement is obtained between the simulated and the empirical results as shown in Figure 4.10, which validates the empirical formula within the limits of the simulated problem. As stated earlier, completely general expression for arbitrary material parameters could be readily found using the same methodology.

A few important observations may be made. First, Q more than doubles (from 400 to over 800) by progressive removal of the dielectric filling, which is very substantial increase. Second, except for the highest values of Q (obtained with nearly maximum sized air cavity), there is practically no dependency on cavity depth d . This means that

there is no reason to compromise the structural rigidity of the LTCC module by removing the maximum number of dielectric layers for large Q . Finally, use of Equation (4.16) for Q in combination with equation (4.15) for f_o or equation (4.11) for ϵ_{eff} allows the designer to rapidly assess the basic performance of VWG cylindrical resonators without any EM CAD tools.

Figure 4.10. Simulated first-mode Q (dots) and the associated fitting curves (lines) versus air volume parameters (A and d).

4.4 Excitation discussion

The preceding analysis of VWG cylindrical resonators did not take into account the feeding mechanism, as the purpose was to investigate the natural resonance behavior of the proposed structures. In Chapter 3, at least three practical ways were seen to exist to excite the resonators, namely CPW strip, probe and loop feeding. In this section, probe and CPW strip excitation will be studied in further detail for VWG cylindrical resonators.

In essence, the goal is to assess which modes may (or may not) be excited by a certain choice of feed type and location, and what will be the resulting resonant frequency shift due to this choice. A baseline resonator design having $d_1 = 7$ mm, $A = 4$ mm, and $d = 5/7$ is employed for the study, which is representative of the VWG structures in this work.

The various probe locations to be analyzed are illustrated in Figure 4.13. These locations are P1 (centre of the dielectric region), P3 (centre of the air region), and P2 (midway between the two).

Similarly, the CPW feed mechanism is shown in Figure 4.14 with identical definitions of P1, P2 and P3. The main difference compared to probe feeding is that the feed line is buried and passes through the narrow wall of the VWG resonator, requiring a separate cavity (to the left of resonator in the figure) for measurement probe access.

For these structures, a modal s-parameter analysis is performed and the resonance frequencies are determined from dips in the S_{11} curves. The results of the various excitation cases are summarized in Table 4.1.

Figure 4.11. one-layer and two-layer probe depth (for half structure).

Figure 4.12. A 3D model for the first probe position (a), with a zoom into the probe region (b) to show the port, the CPW feed and the probe via going through an aperture in the ground plane .

Figure 4.13. Top view for the cylindrical resonator showing the three different probe positions under test.

Figure 4.14. CPW feeding of cylindrical VWG resonator

Table 4.1
Comparison of Probe and CPW excitations

Eigen Mode		Probe Excited						CPW Excited					
		P ₁		P ₂		P ₃		CPW ₁		CPW ₂		CPW ₃	
		f ₀	Δf_0 %	f ₀	Δf_0 %	f ₀	Δf_0 %	f ₀	Δf_0 %	f ₀	Δf_0 %	f ₀	Δf_0 %
f ₀₁	22.11	22.10	-0.05	22.15	0.23	22.15*	0.23	-----	-----	22.10	-0.05	22.15	0.23
f ₀₂	29.00	29.15	0.52	-----	-----	-----	-----	-----	-----	29.05	0.17	29.01	0.03
f ₀₃	32.89	33.30	1.25	-----	-----	-----	-----	-----	-----	33.20	-0.30	33.10	-0.60

For each excitation method and location, the resonant frequencies of the first three modes are listed and compared to the natural resonant frequencies obtained from eigen-mode analysis. Where no frequency is indicated, the desired mode failed to be excited. For the others, the frequency shift is seen to range from less than 0.1% to 1.2%. The corresponding Q values are not shown, but there variation is minimal.

Although position P1 is optimum (in terms of minimum frequency shift) for probe feeding, it is the worst for CPW feeding because none of the modes is excited. It should be noted that position P3 can only be used with probe excitation if the probe is sufficiently deep (in this case, 2 layers deep, marked with a star in the table). The optimum location for CPW feeding is P2 since the first mode frequency shift is minimum and the strip length is shorter than for P3. This will lead to a slightly higher Q.

4.5 Design and Implementation of A Novel Cylindrical High-Q LTCC Resonator

Having previously studied VWG cylindrical resonators in considerable detail, it is now possible to exploit these findings in an optimum design. This section presents the practical implementation of a very high Q LTCC cylindrical resonator.

4.5.1 An optimized design for high-Q

As done in Chapter 3, the following techniques are employed to achieve the highest possible cylindrical resonator Q: the lowest possible loss LTCC ceramic material is selected, using largest possible air filling, employing staggered double-via walls, and choosing the excitation method appropriately [68]. The optimized design carried out here consists of a probe-fed cylindrical resonator with an inserted air cavity and double via walls, as shown in Figure 4.2 [69]. Seven layers out of eight are used to build up the resonator height and only one layer is left for the probe excitation. Other resonator dimensions are shown in Table 4.2.

The spacing between the vias within the via-walls is minimized by using the lowest possible pitch (400 μm , which is less than one tenth of the guided wavelength), in order to eliminate the radiation losses. On each LTCC layer, the vias are connected together through a ring strip making the via wall look like a grid wall so as to ensure good conductivity between all grounded vias, the metallic base and the metallic cover.

The probe is placed close to the center of layer one such that it excites an electric field maximum in the air-filled region of the resonator. Its depth is chosen to obtain appropriate coupling at the desired TM_{010} resonant mode.

4.5.2 Experimental results

The cylindrical design is fabricated using the LTCC tape system described earlier. Figure 4.15 shows a photograph of the fabricated module. The optimized dimensions as well as the measured data are shown in Table 4.2. The unloaded Q is extracted from the measured S-parameters of Figure 4.16 using the techniques described in [71],[72],[73]. The best Q_0 was found to be 840 for the first resonant mode, which is a very high value for a standard LTCC process. In addition, its size is considerably smaller than would be possible with previously reported techniques [68], [16],[58],[61].

Table 4.2
Cylindrical VWG High-Q LTCC resonator

Diameter [μm]	Air Cavity size	Mode	Predicted		Measured		Error %	
			f_0 [GHz]	Q_0	f_0 [GHz]	Q_0	Δf_0	ΔQ_0
Inner:7000 Outer:7600	4150x 4150 [μm^2]	1st	23.75	800	23.96	840	0.88	4.8
		2nd	31.65	290	32.19	281	1.6	-3.5
		3rd	37.5	310	37.3	326	-0.54	4.9

Figure 4.15. Photograph of the cylindrical resonator of Figure 4.2.

As can be seen, agreement between measured and experimental results is excellent, with errors less than 2% for frequency and less than 5% for Q. These errors are easily explained by the accumulative effects of LTCC dimensional tolerances (on the order of +/- 0.7% per feature). Incidentally, considerable sub-banding is required to achieve adequate frequency resolution from the network analyzer, in order to extract accurate Q values from measurements. On the other hand, not all the simulated S_{11} dips refer to a physical resonance mode. Therefore, the field distribution at each dip should be studied and compared with theory. Doing so, the two small minima at 28 and 29 GHz are found to be spurious modes due to accumulated simulation errors.

Figure 4.16. A comparison between measured and simulated return loss for the novel cylindrical resonator of Figure 4.15.

Figure 4.17 shows that a maximum field density exists for the intended mode mainly within the air space. This is in sharp contrast to the field distributions of the possible higher modes, which are primarily located in the surrounding dielectric space.

The lower Q values associated with modes 2 and 3 in Table 4.2 are therefore expected. Clearly, tradeoffs between Q factor, operating mode, air volume, resonator size and fabrication limitations must be made to produce optimum resonators.

(a) Electric field distribution of the first resonant mode

(b) Electric field distribution of the second resonant mode

(c) Electric field distribution of the third resonant mode

Figure 4.17 Comparison of the position of the maximum electric field for each mode.

4.6 Conclusion

In this chapter, MMW cylindrical LTCC resonators have been investigated for the first time. Simple analytical expressions have been derived which relate the VWG heterogeneous structures to their conventional cylindrical resonator components through the use of μ_{eff} and ϵ_{eff} . This allows the designer to rapidly the basic performance of these novel structures without resorting to any 3D EM CAD tools. A parametric analysis has been carried out showing the effects of the available design parameters on resonant frequency and Q. In addition, probe and CPW strip feeding techniques have been investigated in terms their ability to excite desired modes. The above investigations, which complete the third thesis objective, represent a considerable contribution to the field of MMW components and packaging.

Finally a novel LTCC cylindrical resonator was designed, optimized and fabricated, yielding a measured unloaded Q of 840 at K-band. This value is very high for a standard LTCC process. Further, cylindrical resonators offer an almost 50% size reduction compared to corresponding rectangular LTCC resonators.

The novel high-Q resonator of this chapter will be utilized to enhance a K-band oscillator, as will be shown in the next chapter.

CHAPTER 5 Proposed Low-Phase Noise K-/Ka-Band Oscillators Utilizing High-Q Resonators

5.1 Introduction

High capacity digital communication systems require high-quality, low-cost frequency sources. Digital radio links are sensitive to phase noise, which is detected along with the desired signal; the effect is a degradation of the system's bit error rate performance. The final objective of this thesis is to enhance the phase noise, reliability and repeatability of millimeter wave oscillators by proposing an LTCC compatible high-Q resonator and MMIC active circuit combination. In this chapter, low phase-noise oscillators of this type are introduced as a potential solution to the problems of both hybrid waveguide / dielectric-resonator (DR) oscillators and fully monolithic (planar-resonator) oscillators, [75]. The proposed oscillators make use of the novel rectangular resonator of chapter 3 and the novel cylindrical resonator of Chapter 4.

This chapter consists of 7 sections. Section 5.2 reviews the recently reported K/Ka-band, low noise oscillators. Section 5.3 highlights some important theoretical models for MW/MMW oscillators, while Section 5.4 briefly reviews the phase-noise of microwave oscillators and a few techniques to minimize it. In Section 5.5, the active part of a low phase-noise oscillator is designed, optimized, fabricated and measured. Based on this MMIC active circuit, two K/Ka-band oscillators are then realized with the LTCC resonators. Finally, Section 5.6 verifies the predicted results throughout the measurement and characterization of the integrated oscillator modules.

5.2 Review of Recently Reported K-/Ka-band oscillators

A large number of K/Ka-band oscillator designs have been reported over the last few years, and they cannot all be discussed in this work. However, the most relevant representative realizations in the past four years are briefly summarized here. The examples are grouped into the two main categories of free-running oscillators and oscillators employing some type of locking mechanism. In either category, monolithic (single-chip) and hybrid implementations may be found. Table 5.1 presents their major features in tabular form.

In terms of purely monolithic free running oscillators, [76] used an InGaAs HBT technology to achieve at 31.7 GHz, a phase noise of -87 dBc/Hz at 100 KHz offset. Meanwhile [77] reported -78 dBc/Hz at 100 KHz for a 30 GHz oscillator design using a $0.2\mu\text{m}$ GaAs PHEMT process. The lower phase noise encountered in the HBT design may be attributable to low flicker noise of HBT devices, but the flicker noise up conversion mechanism and the resonator Q could alter this result.

Still better phase noise has been demonstrated in HBT technology using a common-emitter (CE) topology rather than common-base (CB) as shown by [4], who achieved -91 dBc/Hz at 100 KHz offset for 36 GHz. However, the output power for this design is not mentioned, indicating it may be quite low. A differential HBT oscillator topology has been shown to provide wide tuning range [5], but also with low output power. For MMIC Ka-band oscillators, PHEMT technology generally offers the largest output power as exemplified by the 11.5 dBm, 21 GHz oscillator described in [78]. By means of comparison, SiGe MMIC technology has been used at K-band, although with substantially poorer phase noise and output power performance [6].

Table 5.1

Recently reported K/Ka-band oscillators.

Ref.	fo	Po	Oscillator SSB phase noise				Hybrid/ Monolithic	Comment
			Free-running Osc.		Locked Osc.			
			at 100KHz	at 1MHz	at 100KHz	at 1MHz		
[76]	31.7	2.6	-87				M	HBT
[77]	30	10	-78				M	PHEMT
[77]	24.4	6	-89				M	PHEMT
[4]	36		-91				M	HBT
[5]	20	-3		-87.6			M	HBT
[78]	20.8	11.5	-82				M	PHEMT
[6]	25	-11	-70				M	SiGe
[79]	36.8	16	-100				H	Gunn / Cyl. Cavity
[74]	26.17	12.6		-118.7			H	MESFET DRO
[80]	35.8	2				-104	H	quadruple / ring reson.
[81]	19	2		-120			H	Hybrid
[82]	19.5	10			-100		M	MESFET / balanced
[83]	24	4.4			-97		H	phase locked / hair pin
[7]	23	0			-150		H	PLL / cryogenic sapphire

Hybrid K/Ka band integrated oscillators can provide substantially lower phase noise than monolithic designs, due to their use of higher quality resonators. Ruimi has demonstrated a 37 GHz oscillator with -100 dBc/Hz phase noise at 100 KHz offset using a Gunn diode in a cylindrical waveguide cavity resonator [79]. A more compact DRO realization has been described in [74], reporting -119 dBc/Hz at 1 MHz offset for a 26 GHz oscillator. Both designs exhibit output powers well above 10 dBm. A novel 36 GHz quadruple-push / ring resonator oscillator topology featuring simple assembly has been presented in [80] using Teflon substrate and discrete components, while [81] describes that the performance of tuning circuit can be improved by ensuring its reflection coefficient has a steep phase slope.

An alternative way of obtaining low phase noise oscillators in either monolithic or hybrid technologies is to employ some form of locking. [9] demonstrates a balanced PHEMT MMIC oscillator at 19.5 GHz with -100 dBc/Hz at 100 KHz, using injection

locking to a synthesized source. Also, a phase-locked hair-pin oscillator can be used to enhance the characteristics of harmonic oscillators, as in [83].

Perhaps the ultimate in low phase noise performance is the design presented in [7], which employs a phase-locked loop (PLL) with cryogenic sapphire resonator to achieve -150 dBc/Hz at 100 KHz offset at 23 GHz. Nonetheless, the requirements for future generations of compact mobile low cost mm-wave communication terminals will likely call for simpler implementations. As this summary has shown, an important gap remains in the current state of the art for filling these needs with simple free-running oscillator designs. The following sections pursue an attractive way of realizing low-phase noise, compact size, repeatable, and low cost free running oscillators.

5.3 Microwave Oscillator Design Principles

Generally speaking, oscillator design has evolved along three parallel paths. The first and most widely used approach is to achieve conditions under which oscillations can start. This will be more fully explained throughout this section, as it applies to the designs of this chapter. The second approach applies van der Pol ideas to develop microwave oscillators. The third uses van der Pol ideas to characterize nonlinear oscillations. The second and third approaches are seldom employed by microwave engineers because the cubic nonlinearity of van der Pol equation cannot be applied to transistor oscillators. However, van der Pol ideas have given rise to the theory of nonlinear oscillations, periodically forced relaxation oscillations and the dynamics of feedback systems [84]. Therefore, microwave researchers may revisit Van der Pol theory to investigate more useful concepts.

The several variants of the first approach to oscillator design may be reduced to two main models. The first is the positive feedback closed loop stability analysis model. The second is the negative resistance model. Both will be described below.

5.3.1 Positive Feedback Analysis Model

An oscillator can be modeled as an amplifier with a positive feed back frequency selective network as shown in Figure 5.1. According to classical oscillator theory, the output frequency is the frequency that satisfies the Nyquist criteria [85]:

$$H(S) = A(s).B(s) = 1 \quad (5.1)$$

Figure 5.1Positive feedback oscillatory system [85].

This relation is valid for the steady-state condition and could be rewritten as:

$$|A(j\omega)| \cdot |B(j\omega)| = 1 \quad (5.2)$$

$$\text{Arg}[A(j\omega)] + \text{Arg}[B(j\omega)] = 2n\pi \quad (5.3)$$

where $n = 0, 1, 2, \dots$

$A(s)$ and $B(s)$ need to be calculated when the amplifier and the filter each are loaded by the filter and the amplifier, respectively. When this condition is satisfied, $A(s) \bullet B(s)$ is calculated by breaking the loop at a plane where the value of the impedance presented by the circuit is well known. Usually this point is the input of the amplifier. For

unilateral amplifiers ($S_{12} = 0$), the input impedance can be obtained from the reflection coefficient S_{11} only. But in some practical cases, S_{12} is not equal to zero, and therefore, the input impedance of the amplifier will be influenced by the load of the amplifier.

The above conditions are based on the Nyquist criteria. They imply that any feedback system can oscillate if its loop gain and phase shift are chosen properly. In most microwave oscillators, however, a frequency selective network (resonator) is included in the loop so as to stabilize the frequency, although such a network can be a part of $A(s)$ as well.

An important advantage of the gain-loop method is its capability to control the output frequency of the oscillator with the external feedback network, and to avoid the effects of unstable amplifiers. This control over the stability of the amplifier is not easily obtained with other analysis methods, such as the negative resistance model. Another advantage is its usefulness in evaluating the noise behavior of the circuit. The higher the frequency, the more difficult to use the gain loop method due to a higher S_{12} , which is the main disadvantage of this method [86].

5.3.2 Negative Resistance (Reflection) Model

The negative resistance method divides the oscillator into two parts: the negative resistance generator and the resonator [87], as shown in Figure 5.2. A noise source is included to help understanding Section 5.3.1. The classic oscillator theory of negative resistance establishes that the impedance presented by the resonator must be equal to the impedance presented by the negative resistance generator (with opposite sign) to begin oscillations. The equations

$$\operatorname{Re}(Z_c) = -\operatorname{Re}(Z_d) \quad (5.4-a)$$

$$\operatorname{Re}(Y_c) = -\operatorname{Re}(Y_d) \quad (5.4-b)$$

are satisfied in any place in the circuit at the oscillation frequency in the steady-state condition. Where:

$Z_c(\omega)$ is the impedance looking into the resonant circuit from the active device.

Z_d : is the impedance of the active device.

$$Z_d = -R_d + j X_d \quad (5.5)$$

Figure 5.2. An equivalent circuit of a one-port negative-resistance oscillator [85].

This method is based on maximizing the negative resistance of the active network [88]. For design and optimization, it is often more convenient to obtain these quantities in terms of the reflection coefficients as follows:

$$\Gamma_d = \frac{Z_d - Z_o}{Z_d + Z_o} \quad (5.6)$$

$$\Gamma_c = \frac{Z_c - Z_o}{Z_c + Z_o} \quad (5.7)$$

Substituting by Eq. 5.6 and 5.7 in Eq.5.4a we get:

$$\Gamma_c = \frac{1}{\Gamma_d} \quad (5.8)$$

Which is the steady state one-port oscillation condition in terms of reflection coefficients. For a passive circuit, which is the common case, the magnitude of Γ_c will be between 0 and 1. Thus, in order to get oscillation, Γ_d must have a magnitude larger than one.

If the circuit has high-Q, then it varies in frequency much faster than the active part of the circuit. Consequently, Γ_d can be considered independent of frequency, and the criterion can be rephrased in a form particularly convenient for applications: The system is unstable if the contour of Γ_c encircles $1/\Gamma_d$. Consequently, it is easy to change parameters of the active circuit so that the encirclement happens.

For the case of microwave transistor oscillator, employing an N-port oscillation condition is more beneficial. Such techniques are well described in details in standard textbooks [89], [90]. Also, stability considerations lead to check two more factors; namely the inherent stability factor of a two-port network $K(S)$ [91] and the Mu-stability parameter $Mu(S)$ [92].

5.3.3 Phase Noise in Microwave Oscillators

As with all circuits, oscillators are susceptible to noise. A real oscillator produces noise in the form of random phase and amplitude distortions that affect the carrier signal. In transmitters, oscillator noise can result in adjacent channel interference and modulation errors. In receivers, oscillator noise can result in demodulation errors, reduced sensitivity,

limited dynamic range and increased bit error rate. Noise injected into an oscillator by its constituent devices or by external means can influence both the frequency and the amplitude of the output signal. In most cases, the disturbance in the amplitude is negligible or unimportant and only the random deviation of the frequency is considered. The latter can also be viewed as random variation in the period or deviation of the zero-crossing points from their ideal position along the time axis. For a nominally periodic sinusoidal signal, we can write,

$$x(t) = A \cos[\omega_o t + \phi(t)] \quad (5.9)$$

Where $\phi(t)$ is a small random excess phase representing variations in the period. The function $\phi(t)$ is called “Phase Noise”. Note that for $|\phi(t)| \ll 1$ rad, we have,

$$x(t) \approx A \cos(\omega_o t) - A \phi(t) \sin(\omega_o t) \quad (5.10)$$

That is the spectrum of $\phi(t)$ is translated to $\pm \omega_o$.

All practical oscillators inherently possess an amplitude-limiting mechanism of some kind. Because amplitude fluctuations are usually greatly attenuated as a result, phase noise generally dominates. In microwave applications, phase noise is usually characterized in the frequency domain. For an ideal sinusoidal oscillator operating at ω_o , the spectrum assumes the shape of an impulse, whereas for an actual oscillator, the spectrum exhibits “skirts” around the carrier frequency, as shown in Figure 5.3. To quantify phase noise, we consider a unit bandwidth at an offset $\Delta\omega$ with respect to ω_o , calculate the noise power in this bandwidth and divide the result by the carrier (average) power.

Figure 5.3. Output spectrum of ideal and actual oscillators [98].

(A) Kurokawa's Model

In 1968, Kurokawa analyzed the noise characteristics of the electromagnetic output of a negative resistance one-port oscillator [93] indicating the AM noise and the phase noise in terms of power spectral density. This noise model is applicable for Gunn and Impatt oscillators. Then it was extended to a two-port oscillator analysis [94]. Daniel and Howes [95], indicates that the phase noise spectrum depends on the vectors tangential to the circuit and device loci in a similar manner to the AM noise. They proposed that a good noise performance might be achieved if the circuit and the inverse device loci of Figure 5.2 intersect orthogonally at ω_o . Then some techniques for designing a microwave oscillator with low phase noise are introduced in [96].

(B) Leeson's Model

Early in 1960s, feedback oscillator was analyzed using a simple model and assuming white noise to be associated with the active device. In 1966, Leeson extended the analysis to include the effects of the flicker ($1/f$) noise and the noise floor [97]. The normalized single-sideband (SSB) phase noise for oscillators, with an added term for flicker noise, (as shown in Figure 5.4) is

$$L(f_m) = 10 \log \left[\left(\frac{FkT}{2P_s} \right) \left\{ 1 + \left(\frac{f_0}{2Q_l f_m} \right)^2 \right\} \left(1 + \frac{f_c}{f_m} \right) \right] \quad (5.11)$$

Where, $\frac{FkT}{P_s}$ is the asymptotic power spectral density of the phase

modulation term $\phi(t)$ which is defined in Eq. 5.16

F = active device thermal noise factor;

k = Boltzmann's constant = 1.38×10^{-23} J/K

T = temperature °K

P_s = amplifier input power level

f_o = oscillator fundamental frequency

f_c = flicker-noise corner frequency (sometimes called f_{1/f^3} or ω_{1/f^3})

f_m = frequency offset from carrier (sometimes named Δf or $\Delta\omega$; if angular)

Q_l = Loaded quality factor of the resonator (corrected by [75])

Figure 5.4. SSB Oscillator phase noise output spectrum [98].

Therefore, the phase noise improves as both the carrier power and Q_l increase. Increasing the signal power improves the ratio simply because the thermal noise is fixed, while increasing Q improves the ratio quadratically. Unfortunately, using active circuits to boost Q , misled the designers because the noise factor F is increased preventing the anticipated improvements in phase noise. Comparing 2 similar oscillators with different resonators, it has been shown that a 10 times higher resonator Q leads to about 10 dB phase noise enhancement [84].

(C) Lee and Hajimiri's Model

Finally, Lee and Hajimiri introduced their linear time variant model [99]. They proved that the time invariance assumption of the other conventional models fails and assumed the oscillator to be a linear, but (periodically) time-varying (LTV) system. It identifies an important symmetry principle, which may be exploited to suppress the up conversion of $1/f$ noise into close-in phase noise. This new model provides a quantitative reconciliation between theory and measurements. The main conclusions of the Lee and Hajimiri's model that are related to our work can be summarized as follows:

1) First, in common with the revelations of LTI models, both the signal power and resonator should be maximized, all other factors held constant.

2) In addition, phase shifts around the loop generally must be arranged so that oscillation occurs at or very near the center frequency of the resonator. This way, there is a maximum attenuation by the resonator of off-center spectral components.

3) The $1/f^3$ phase noise corner is not necessarily the same as the $1/f$ device/circuit noise corner; it will generally be lower. In addition, there is a possibility of reducing by

large factors the $1/f^3$ phase-noise corner. This result has particular significance for technologies with poor $1/f$ noise performance, such as CMOS and GaAs MESFET's.

Although brief, this phase noise discussion will be useful for developing low noise oscillators in the following sections. Implicit throughout is the need for high Q resonators to achieve such a result. Also, choosing the appropriate circuit topology and applying the appropriate nonlinear design and optimization techniques are crucial.

5.4 Development of Novel K/Ka-band Novel LTCC-based oscillators

5.4.1 Design Strategy and target Specifications.

The realization of LTCC-based oscillator modules requires access to both an active device technology and an LTCC fabrication process. In order to obtain the smallest possible oscillator module, it was decided to design an MMIC chip containing all the oscillator circuitry except for the resonator element, which could be embedded within the LTCC package. Since this work was carried out over a period of 3 years, and the access to the individual MMIC and LTCC technologies and their fabrication schedules could not be coordinated, the active MMIC circuit was designed separately from the LTCC resonators. This approach required some compromises to be made in terms of operating frequency, tunability and optimization of output power

For demonstration purposes, oscillators operating in the 20-30 GHz range are to be designed, fabricated and measured. The specifications are shown in Table 5.2, which are loosely based on the requirements of the point-point digital radio systems being developed for the 24 GHz ISM and 29.5 GHz LMDS bands.

The design of active MMIC component for the oscillators will be described in Section 5.4.2 while the integration with the LTCC resonator and the final measurements will be covered in Section 5.4.3

Table 5.2
Specifications of the novel oscillators

Oscillator Specification	Oscillator 1	Oscillator 2
Frequency of Oscillation	24 GHz	29 GHz
Output Power	10 dBm or better	5 dBm or better
Phase Noise	-100 dBc/Hz at 100 KHz offset or better	-100 dBc/Hz at 100 KHz offset or better
DC to RF conversion efficiency	10 % or better	10 % or better
Power Supply	< 3 V	< 3 V
Maximum harmonic/sub-harmonic level	-30 dBc	-25 dBc
MMIC physical chip area	1mm X 2mm	1mm X 2mm
Packaged oscillator size	~1 Cm ²	~1 Cm ²

5.4.2 Design, Simulation and optimization of a wide-band Unstable MMIC active circuit

Monolithic-microwave integrated-circuit (MMIC) designs for millimeter-wave applications may employ MESFET, HEMT or heterojunction bipolar transistor (HBT) technology. Although a GaAs PHEMT technology was chosen for this work based on our foundry access through CRC, it is important to briefly evaluate such a choice in the context of oscillator design. This is because in solid-state microwave oscillators, a considerable amount of phase noise is generally attributed to the active device. The device low frequency noise (i.e. flicker noise) up-converts as noise sidebands at offsets from the RF carrier signal with certain up-conversion factor. Therefore, for a stable oscillator design, it is of utmost significance to consider both the LF noise and the noise up-conversion mechanism of the device. The use of HBT's as the basis of millimeter-wave oscillators has been given much attention recently due to the superior low-

frequency noise characteristics of HBT's. However, the output power available from HBT-based oscillators cannot compete with the HEMT-based oscillators [77], [100],[101],[102]. On the other hand, a lower noise up-conversion factor can be achieved in HEMT than in HBT because of the inherent linear behavior of HEMT over certain bias conditions [103],[104],[105]. On balance, therefore, use of a HEMT device is an appropriate choice for achieving a low phase noise oscillator with relatively large output power.

Specifically, the MMIC process to be employed is a 0.2 μm gate length pseudomorphic HEMT low phase noise technology (ED02AH) from *Philips Microwave Limeil Foundry, France*, on 100 μm GaAs substrates. Refer to Appendix B for further details. Here, we will focus on optimizing an active network to work as an unstable circuit (negative resistance generator) for any oscillator in the range of frequencies of 20 to 30 GHz. Then, any high-Q resonator from Chapter 3 or 4 could be employed to tune the developed hybrid combination at the desired frequency, keeping the low phase noise target in mind.

A circuit topology is selected as shown in Figure 5.5. The common-source configuration is chosen, because of its superior frequency stability due to the high isolation between the frequency determining elements and the oscillator output. However, because of the small value of the transistor's gate-drain capacitance C_{gd} the instability of the circuit has to be achieved using external elements. Therefore, a feedback interstage is designed which leads to an input reflection coefficient (S_{11}) higher than unity. Figure 5.5 has a blocking capacitance C_{RF} and an LR-network which may be used in order to achieve the transistor instability. However, in our case, a series inductive feedback is

chosen since it can be realized by a short-circuited microstrip stub that facilitates the flow of the DC drain current to ground. So, the circuit topology of Figure 5.6 is chosen to create instability over a wide frequency band.

Figure 5.5. Conventional common source configuration.

Figure 5.6. Schematic circuit of the proposed K/Ka-band Oscillator.

The transistor is chosen to be pHEMT 6x50 μm . This transistor ensures an adequate amount of power to overcome the losses in the LTCC resonator and other passive MMIC circuits. Before choosing an appropriate bias circuit, we must select a bias point to determine the required drain current. $I_{DS} = (I_{DSS} / 2)$ is chosen as a bias point because it provides the best compromise among gain, saturated power, and low harmonic output. So, the required current will be 36 mA, at $V_{ds}=3\text{V}$, for the chosen pHEMT in this work.

At the drain, a biasing circuit is designed using a combination of a microstrip thin line and a small inductance as shown in Figure 5.7. The used spiral inductor is designed with the maximum available inner area and the maximum separation between lines to decrease its capacitive effect. Its optimum value (0.59 nH) is simulated separately with a grounded end to make sure that the self-resonance of the coil itself is away from the frequency band of interest. This bias network is optimized to form a very high impedance to avoid the leakage of any microwave signal in the range of 20-30 GHz through the V_{DD} supply. The transmission line should be as narrow as the available technology allows, but due to current capability issue it is adjusted to be 10 μm . This bias circuit is optimized then simulated using Momentum EM simulator to include any coupling effects, ignored in circuit simulation. Figure 5.7 shows the grid placement of the bias circuit layout. The impedance of the bias circuit is almost 1 $\text{K}\Omega$ or larger in the desired band. This guarantees that the bias circuit achieves both high impedance and wide band objectives.

At the gate, a simple $2\text{K}\Omega$ resistance is used with a quarter wavelength high impedance very thin line connected directly to the V_{GG} supply. From ac point of view this microstrip line is grounded through a capacitor. The schematic circuit of the proposed

oscillator active network is shown in Figure 5.8. This figure shows the pHEMT transistor, the feedback network, the gate and drain bias networks, the coupling capacitors, and any other layout artwork needed to connect all of them.

Figure 5.7. Drain bias circuit (showing the grid for momentum EM simulation)

An optimization routine is performed to ensure the instability of the obtained active network of Figure 5.8 in the frequency range from 20 to 30 GHz, by adjusting the lengths of L_g , L_1 , L_2 and L_{fb} stubs. The optimized active network is simulated using small-signal S-parameters from 10 to 30 GHz. Figure 5.10 shows that the circuit became unstable since:

$$Mu1 < 1 \quad \text{and} \quad |S_{11}| > 1$$

This means that port one (the gate side) is generating negative impedance if port 2 (the drain side) is terminated by 50 Ohm. Because the oscillation phenomena inherently depend on the nonlinear operation of the transistor, the large-signal S-parameters are simulated as well, giving almost the same instability from 15 to 30 GHz.

So far no resonator has been employed in the design. In order to choose the appropriate resonator, it should be in the unstable region of the smith chart. For this reason, the input and output stability circles are simulated at 20, 25, and 30 GHz as shown in Figure 5.9. Since $|S_{11}| > 1$, the unstable region for each frequency is the side at which the corresponding circle encircles the origin.

Figure 5.10. Input and output reflection coefficients and MU stability factors of the MMIC active network

Figure 5.9. Input and output stability circles of the MMIC active network

5.4.3 Simulation results of the proposed 24GHz oscillator

For the sake of completing the simulation of the proposed oscillators, an ideal 24 GHz resonator in the form of RLC model is employed to check the oscillation conditions. The design procedure of this section follows closely the improved nonlinear analysis and design method of [106]. Here, the small signal loop gain test as well as the harmonic balance nonlinear simulation will be performed.

(A) Small-signal S-parameter loop gain test

The feedback loop is broken and each part of the circuit is analyzed separately. If the loop gain is greater than one with a phase shift of zero, the circuit may oscillate. A special oscillator test element is provided to check for this condition, as shown in Figure 5.11. The Nyquist condition is sought by the oscillator test component "OSCTESTG", as:

$$S_{ii} = \frac{b_i}{a_i} = 1 \angle 0^\circ ; \quad \text{where } a_i \text{ and } b_i \text{ are as indicated in Figure 5.11.}$$

Figure 5.11. A test port is braking the loop to check for oscillation conditions.

ADS uses an iterative process to try to find a set of non-zero steady-state conditions which indicate oscillation. If the circuit were in a state of oscillation, the loop gain would be exactly one at the fundamental frequency of oscillation. The oscillator port

is inserted into the circuit to break the loop. The oscillator port is actually a special sort of current source. If the current through the port matches the current of the oscillator, the voltage drop across the port is zero.

So, the Computer Aided Analysis of an oscillator proceeds as follows:

- Sweeping the frequency to determine the magnitude of S_{11} at each value. The first oscillation guess is obtained based on the highest magnitude then entered on the analysis block. Subsequent guesses are derived from the results of previous analysis.
- Calculating the voltage drop across the oscillator port. If the voltage drop is zero, the analysis is finished. If the phase is zero, the magnitude of the current source is changed. If the phase is not zero, the frequency of the current source is changed.
- Going back to step one.

If there is no voltage drop across the current source, then the current source supplies no power to the circuit. So, the signal must be coming from the oscillator, not the current source, which is the oscillation that we are looking for.

The small-signal loop gain for our 24GHz oscillator circuit is configured. The loop gain S_{11} has both magnitude and phase. ***The results of the loop gain test are presented on two plots*** as shown in Figure 5.12.

1. A rectangular plot shows the magnitude and phase of the loop gain S_{11} .

If $\text{MAG}(S_{11}) > 1$ when $\text{PHASE}(S_{11}) = 0$, the circuit is likely to oscillate at the frequency where the phase is zero. If $|\text{Loop Gain}| < 1$, the circuit will not oscillate.

Oscillations will build to higher signal levels, where the gain will compress until $|\text{Loop Gain}| = 1$ for large-signal operation which is the steady-state case as will be shown later.

2. **A polar plot** that shows S_{11} in complex form. The trace should encircle the point $(0+j1)$, with increasing the frequency in clockwise direction. It is clear from Figure 5.12, that the circuit will oscillate at 24GHz frequency.

Although this checked condition indicates oscillation start up, at some level the amplitude of oscillation is limited. This limit may come about through the saturation of the active device, or may involve an active gain control network. Either way the limiting action requires that the gain of the loop is a function of the amplitude of the signal. This requires a nonlinear analysis, as will be shown next.

(B) Harmonic balance nonlinear analysis

In this section, the *harmonic balance* technique is used to analyze the harmonic levels of the oscillator, output power and the time domain wave-form of the output voltage. A transparent probe which is called OSCPORTG is inserted in the same position as the OSCTEST port and has the same operation principle except using Harmonic Balance instead of S-parameter simulation. The analysis results are shown in Figure 5.13. The fundamental frequency is exactly 23.98GHz with 16.4 dBm output power. The second harmonic is 23.4dB lower than the fundamental. The third harmonic is 28.4dB lower than the fundamental. Those values are considered to be good in terms of harmonic distortion. Finally, the steady state time domain wave-form is shown in Figure 5.14.

Figure 5.12. Results of loop gain test.

Figure 5.13. Results of harmonic balance simulation.

Figure 5.14. Expected time domain waveform of the 24GHz oscillator.

Similar simulations are performed for the 29 GHz proposed oscillator based on an external RLC resonator model [107]. The simulated results for both oscillators are summarized in Table 5.3.

Table 5.3
Simulated Results of the proposed oscillators

Oscillator Specification	24 GHz Oscillator	29 GHz Oscillator
Frequency of Oscillation	23.98 GHz	29.1 GHz
Output Power	16.4 dBm	10.5 dBm
Phase Noise	-----*	-----*
DC to RF conversion efficiency	10 %	8 %
2nd- harmonic level	-23 dBc	-25 dBc

* Phase noise predictions are not reported here because the available pHEMT models are inadequate, omitting the flicker noise model.

5.4.4 Fabrication and Measurements of the MMIC active circuit

The circuit of Section 0 is fabricated with a chip area of almost $1 \times 2 \text{ mm}^2$, as shown in Figure 5.15. The fabricated circuit is measured on wafer. A probe station is employed for the MIC measurement to secure the chip in place and to provide a stable means of electrically connecting it to the measurement apparatus. A network analyzer was used to perform the S-parameter measurements. The measurement apparatus consisted of an HP8510XF network analyzer, cabling and probes. The network analyzer

is capable of accurately extracting the S-Parameters of the device over the 45 MHz to 65 GHz frequency band. The HP8510XF was calibrated up to the probe tips with an OSLT (Open, Short, Load, Through) calibration method. Coplanar waveguide probes with V-connectors were utilized for the RF signals. Such probes provide extremely good 50Ω characteristic impedance up to the contacting tips, thus minimizing parasitic reflections and mismatching to the device under test (DUT). In addition, 2 six-pin DC-probes with two power supplies are employed for DC bias. The test setup is shown in Figure 5.16.

The magnitude of S_{11} of the active circuit are plotted for different chips as shown in Figure 5.17. All of them are close to each other and agree with what is expected from simulations. This also proves that the design is repeatable and is not sensitive to the MMIC fabrication tolerances and meets the objective of active elements for 20-30 GHz oscillator.

Figure 5.15. Chip micro photo of the fabricated 20GHz oscillator active circuit.

Figure 5.16. Test setup for S-parameters measurement.

Figure 5.17. A comparison between simulated and measured S parameters from different chip reticules.

5.5 Integration and Measurements of the proposed K/Ka-band oscillators

This section focuses on integrating and measuring two novel oscillator prototypes. The first is the 24 GHz oscillator, which makes use of the cylindrical high-Q LTCC resonator of Chapter 4. The second is the 29 GHz oscillator which makes use of the rectangular high-Q resonator of Chapter 3. Both employ the same MMIC chip of Section 5.5.1 as the oscillator active part.

5.5.1 Integration of the proposed oscillators

Placement of the MMIC chip in the LTCC resonator Package can be done using either bonding or flip-chip mounting. In both cases, the chip is placed on the top LTCC surface for simplicity. These two methods of hybrid integration along with some practical concerns are discussed briefly as follows:

(A) Bonding of the MMIC chip with the LTCC resonator package

The bonding technique employs either ribbons or bond-wires for the electrical signal connections and is a simpler assembly method than flip-chip techniques. However, it is prone to radiation losses which may degrade the achieved high-Q of the LTCC resonators. Based on its availability this method was initially tried. The bonding is done using gold ribbon instead of wires to minimize the parasitic inductance effects. A ribbon width is chosen to be 75 μm , which is smaller than the bond pad width in order to minimize the impedance step discontinuity. The resulting dimensions of the ribbon bonds are: Width = 75 μm , Length = 125:200 μm , and a thickness of 12 μm .

Figure 5.18 shows a bonded chip sample. Measurements of these oscillator modules were performed but found to suffer from parasitic effects. The results of the flip chip mounted oscillators will be presented instead.

Figure 5.18. Micro photo of the ribbon bonded MMIC die placed on the LTCC package just above the resonator port.

(B) Flip-chip mounting of the MMIC circuit on the LTCC resonator package

The flip chip technique consists of mounting the MMIC upside down with the electrical contacts being made through solder bumps. Because the MMIC die was fabricated without any solder bumps, regular solder flip chipping could not be done. Instead, Au stud bumps were used, as these may be formed on the LTCC carrier. Then, the connection is made with the use of a conductive adhesive. Photo 1 of Figure 5.19 shows adhesive bumps before flattening and photo 2 after flattening, It is seen that the flattening is successful since the bumps do not spill over to adjacent pads. Finally, the desired flip-chip mounted oscillator sample is shown in Figure 5.20 with overall dimensions of 11mm x 9.5 mm.

(a) Photo 1

(b) Photo 2

**Figure 5.19. Micro photos of the MMIC die with adhesive bumps
(a) before flattening (b) after flattening.**

Figure 5.20. Photo of the fabricated flip-chip oscillator.

5.5.2 Measurements and characterization of the proposed oscillators

Having successfully obtained compact sized, packaged, LTCC oscillator modules, the goal of this section to assess their performance experimentally. Two oscillators are demonstrated: the 24 GHz design based on the cylindrical resonator of Chapter 4 and the 29 GHz design based on the rectangular resonator of Chapter 3.

(A) Experimental results of the proposed 24 GHz oscillator

The oscillator module was measured on a probing station connected to DC supplies and an Agilent PSA series spectrum analyzer (E448A). Appendix D provides additional details about the probes and cables in this setup. Figure 5.21 gives the oscillator measured output power spectrum, showing the desired 10.3 dBm signal at 24.7 GHz and the second harmonic at -18 dBc. Figure 5.22 plots the single side band phase noise performance for the 24.7 GHz signal. After averaging at a spot frequency of 100 KHz offset, the phase noise level is determined to be -106 dBc/Hz. To complete the characterization, the output power and frequency variations versus gate bias voltage are plotted in Figure 5.23. The DC_RF conversion efficiency for this oscillator is 9%.

Figure 5.21. Measured spectrum for the 24 GHz proposed oscillator.

Figure 5.22. Measured SSB phase noise for the 24 GHz proposed oscillator.

Figure 5.23. Measured Power and frequency variation versus gate bias tuning of the 24 GHz oscillator.

(B) Experimental results of the proposed 29 GHz oscillator

The 29 GHz oscillator is measured in a similar fashion. Its output power spectrum is given in Figure 5.24. The single side band phase noise is shown in Figure 5.25 and the frequency/power versus gate voltage characteristic is plotted in Figure 5.26. The DC_RF

power conversion efficiency of this circuit is 6.5%. Table 5.4 summarizes the main results for the two oscillators, along with the deviations from the predicted values of Table 5.3.

Figure 5.24. Measured spectrum for the 29 GHz proposed oscillator.

Figure 5.25. Measured single side band phase noise for the 29 GHz proposed oscillator.

Figure 5.26. Measured Power and frequency variation versus gate bias tuning of the 29 GHz oscillator.

Table 5.4

Measured Results of the proposed oscillators

Oscillator Specification	24 GHz Oscillator			29 GHz Oscillator		
	Predict.	Meas.	Error %	Predict.	Meas.	Error %
Frequency of Oscillation [GHz]	23.98	24.7	3	29.1	28.9	-0.7
Output Power [dBm]	16.4	13.3	-18.9	13	11.5	-11.5
Phase Noise	-----	-106		-----	-101.5	
DC to RF conversion efficiency	10 %	9%	-10	8 %	6.5%	-18.8
2nd harmonic level from carrier [dBc]	-23	-18	21.7	-25	-----	
Tuning range (MHz)	-----	~160		-----	~100	

Clearly, extremely low phase noise performance has been observed as a result of employing the high-Q resonators within the same LTCC package. In addition, the output powers achieved by these designs are remarkably high. Agreement with the predicted performance is excellent, with only 0.3% to 1.3 % shift in frequency and about 3 dB discrepancy in output power. This confirms the validity of the design method followed in Sec. 5.4. The 100 MHz –160 MHz tuning range is narrow, as expected from simple gate bias control, but broad tuning was never the intent of this study. Appropriate low-loss tuning elements may be employed in the future to obtain ultra low phase noise VCOs.

It is interesting to compare the main performance parameters of the LTCC- based oscillators proposed here with the other reported designs discussed in Table 5.1. Figure 5.27 and Figure 5.28 show these comparisons for phase noise and output power, respectively, confirming that the proposed designs are among the best recently published oscillators.

Figure 5.27. SSB phase noise of the proposed and recently reported oscillators.

Figure 5.28. Output power comparison of the proposed and recently reported oscillators.

5.6 Conclusion

In this chapter, high performance MMW LTCC-based oscillators have been demonstrated for the first time. The novel design, consisting of embedding the high Q resonator in the same LTCC package that contains the active chip, was successfully implemented. A 24 GHz oscillator and a 29 GHz oscillator, utilizing the highest Q LTCC resonators from the previous chapters, produced output powers of 13.3 dBm and 11.5 dBm, and extremely low phase noise of -106 and -101 dBc/Hz. These oscillators are considered to be among the best low phase noise, medium power free running oscillators. The excellent agreement between predicted and measured results proves the validity of the design technique and methodology. This fulfills the final objective of this thesis. It is believed that the proposed oscillators will be very useful in the design of miniaturized, low-cost, mass-producible, low phase-noise MW/MMW systems, since no extra foundry or special processing is needed apart from regular MMIC chip packaging.

CHAPTER 6 Conclusions, Contributions, and Future Work

6.1 Conclusions

High-Q LTCC resonators have been investigated for MMW applications. Our modeling and simulation methodology has been experimentally validated through many different LTCC resonator designs at MMW frequencies. A design strategy for obtaining high Q-factors has been proposed, which includes the insertion of air cavities and novel excitation structures along with design guidelines. Based on these techniques, three novel LTCC high-Q resonators were designed, optimized and fabricated. Measured unloaded Q of 1100, 700 and 550 were obtained at Ka band, for probe-fed rectangular and different loop-fed rectangular configurations, respectively. These values are the highest ever reported for a standard LTCC process and fulfill the first two thesis objectives.

Also, MMW cylindrical LTCC resonators have been investigated for the first time. Simple analytical expressions have been derived which relate the VWG heterogeneous structures to their conventional cylindrical resonator components through the use of a_{eff} and ϵ_{eff} . This allows the designer to rapidly assess the basic performance of these novel structures without resorting to any 3D EM CAD tools. A parametric analysis has been carried out showing the effects of the available design parameters on resonant frequency and Q. In addition, probe and CPW strip feeding techniques have been investigated in terms of their ability to excite desired modes. The above investigations,

complete the third thesis objective and represent a considerable contribution to the field of MMW components and packaging.

In addition, a novel LTCC cylindrical resonator was designed, optimized and fabricated, yielding a measured unloaded Q of 840 at K-band. This value is very high for a standard LTCC process. It is noted that cylindrical resonators offer an almost 50% size reduction compared to corresponding rectangular LTCC resonators.

Finally, high performance MMW LTCC-based oscillators have been demonstrated for the first time. The novel designs, consisting of embedding the high Q resonator in the same LTCC package that contains the active chip, were successfully implemented. A 24 GHz oscillator and a 29 GHz oscillator, utilizing the highest Q LTCC resonators from the previous chapters, produced output powers of 13.3 dBm and 11.5 dBm, and extremely low phase noise of -106 and -101 dBc/Hz. These oscillators are considered to be among the best low phase noise, medium power free running oscillators. The excellent agreement between predicted and measured results proves the validity of the design technique and methodology. This fulfills the final objective of this thesis.

It is believed that the proposed oscillators will be very useful in the design of miniaturized, low-cost, mass-producible, low phase-noise MW/MMW systems, since no extra foundry or special processing is needed apart from regular MMIC chip packaging.

6.2 Thesis Contributions

This work has provided numerous contributions to the fields of millimetre-wave high-Q resonators, LTCC packaging and K/Ka-band oscillators. These may be summarized as follows:

- Achievement of the highest reported Q in an LTCC-based resonator, with a 29 GHz rectangular via wave guide (VWG) structure.
- Development of new cylindrical VWG resonators with rectangular cavities, including the first analytical treatment of their behavior in terms of material filling ratio and excitation mechanisms.
- Study of novel resonator excitation methods suitable for the LTCC environment, and the formulation of design strategies to achieve high-Q performance, along with guidelines and trade-offs.
- Demonstration of the first high performance Ka-band oscillators using MMIC chips packaged in LTCC modules with embedded resonators.

6.3 Future work

While the present work has been successful in achieving a high performance in a free running oscillator, it is important to examine low-loss tuning elements so as to achieve frequency agility without compromising the high-Q. Specifically, MEMS varactors could be used to tune the high-Q LTCC resonators and provide voltage controlled oscillators with extremely low phase noise and compact size.

On the other hand, further research should be undertaken in the field of current controlled LTCC-based oscillators, especially with the emergence of LTCC ferrite materials. It is thought that the resonant frequencies of LTCC resonators could be changed by application of variable DC magnetic field bias.

Moreover, the development of high-Q LTCC resonators is by no means complete. For example, other geometrical structures should be explored as a means of achieving higher Q values for lower frequency designs that employ high dielectric constant LTCC materials. Furthermore, it is important to investigate the effect of the material parameters on the multilayered cylindrical resonators with rectangular cavities to complement the study of the structural parameters examined in this work. This will hence lead to the evolution of a complete mode chart for the proposed resonators.

Finally, the available LTCC foundry processes still face some uncertainty regarding material properties as well as subtle dimensional tolerances that should be well considered by designers. This will facilitate a yield analysis, thus completing the established research in the field of high Q 3D resonators embedded in LTCC packages.

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Appendix A: Specifications of the used LTCC material

A.1. Ferro A6-S LTCC specification data sheet [118]

A6-S is a low cost Microwave Grade Low Temperature Co-Fire Ceramic Tape alternative to A6-M. A6-S features excellent dielectric properties of our famous A6-M but at a fraction of the cost. A6-S formulated with Ferro's patented Calcium-Boro-Silicate glass is designed to be used for low cost wireless frequency applications between 2.45-77 GHz. A6-S is compatible with our all Gold, and mix metal metallization. A6-S formulated with a newly formulated polymer system gives A6-S tape that can easily be laminated and cut in the green state newly formulated Silver Metallization, employing our novel technology SAMS (Self-Adhering Metallization) was formulated with the express purpose to bring a low cost metallization with the lowest resistance known. The low resistance results in unequal performance in your high frequency application a cost that you can afford. The use of SAMS technology

Typical Fired Physical Properties

Thermal Coefficient of Expansion (25 - 300°C)	8 ppm/°C
Tape Shrinkage	
X,Y	14.9 ± 0.2 %
Z (unlaminated to fired)	24 ± 0.2 %
Fired Density	2.45 g/cm ³
Flexural Strength (3 point bend)	>160 MPa
Thermal Conductivity	2 W/mK
(With Thermal Vias)	>50 W/mK

Note: All physical testing is performed on 10-layer tape laminates with no metallization, and processed according to FERRO's recommended processing guidelines.

Typical Fired Electrical Properties

Dielectric Constant (10 MHz)	5.9 ± 0.15
Dissipation Factor (10 MHz)	<0.2 %
Insertion Loss (10 GHz)	<0.18 dB/in
Bulk Resistivity	>10 ¹² Ω/cm

Note: All electrical testing is performed on parts manufactured according to Ferro recommended process guidelines with silver-based conductor systems.

resulted in a material with excellent solderability, leach resistance and both initial and age adhesion. A6-S comes in 60 and 100 micron at a thickness of 18 cm. The thin tape, 60 micron gives the electrical designer the added degree of freedom to design those high frequency circuits where headroom is critical for the design. In addition A6-S also come in our standard thicknesses of 5 and 10 mil at our standard width of 13 inches.

Compatible Materials Systems

FERRO A6-S should be used with thick film materials from the FERRO A6 LTCC All Silver-Based (ASB), All Gold-Based (AGB) or Mixed Metal Materials (MMM) Systems. See *data sheets for details*.

Typical Processing Guidelines

Lamination 3,000 psi (21 MPa)
 70° C, 5-10 minutes

Burnout 450°C for 2 hours
 ≤2°C/min. ramp to 450°C.

Firing 850°C peak for 10 minutes,
 Ramp to peak temp 6°C-8°C/min.

Setters Silicon Carbide, fused quartz.
 Bare alumina setters should not be used.

See FERRO A6 LTCC Processing Guidelines data sheets
for full details.

Storage Conditions

We recommend storing A6 in a cleanroom environment if at all possible. Temperature should be ~22°C (70°F) and RH ~ 40%. The tape should be kept in the sealed bags it arrived in until ready for use. Resealing after each use, if possible, is a plus.

A.2. Ferro A6-S LTCC Measured specification

The average relative permittivity as well as the loss tangent of Ferro A6-S are taken from the measured data as extracted in [57]. Two ring resonators, Ferro1 at 5.8 GHz and Ferro 2 at 5.2 GHz, were carried out in Ferro K6-S system using the same foundry process at VTT Electronics , Oulu, Finland.

Fig. A.1 shows the measured relative permittivity constant versus frequency as extracted from the two employed ring resonators. With extrema of 6.14 and 6.22, the average relative permittivity of the Ferro system is 6.19, which is 4.9% higher than the nominal value listed for this material system [118].

The loss tangent of Ferro A6-S. is shown in Fig. A.2. Although it is a bit noisy, it seems to have the lowest loss compared to other commercially available LTCC materials if the first point of Ferro 2 resonator is neglected.

Fig.A.1. Relative permittivity of Ferro K6S as measured in [57] using 2 ring resonators.

Fig.A.2. Loss tangent of Ferro K6S as measured in [57] using 2 ring resonators.

Appendix B: Philips OMMIC ED02AH Process

The OMMICED02AH process used in the MMIC fabrication for this thesis is developed by Philips Ltd. for applications up to millimetric frequencies, and for high rate digital circuits for optical links. The transistors are 0.2 gate length depletion mode pHEMTs. In addition, in this process, enhancement mode transistors, diodes, implanted and thin-film resistors, Metal-Insulator-Metal (MIM) capacitors, spiral inductors and air bridges can be made. The ED02AH process provides several backside options. A 620 thickness GaAs substrate without backside metal and via holes is suitable for coplanar design. On the other hand, a 100 thickness GaAs substrate with or without via holes is available for microstrip design.

The ED02AH GaAs pHEMT transistor is based on an AlGaAs-InGaAs-GaAs heterostructure obtained by MOVPE, using an industrial multiwafer system. The cross section of this device is shown in Figure B.1.

Figure B.1: Cross section of ED02AH pHEMT.

The transistors may be laid out with one or two source grounds. Normally, symmetrical grounding is used to obtain the best performance. The small signal equivalent circuit for the device is shown in Figure B.2..

Fig. B.2. sSmall signal equivalent circuit for the pHEMT device.

where:

C_g , C_d , C_s , L_G , L_D , L_S are the external parasitics of the pHEMT.

R_g , R_s , R_d are the pHEMT access resistances.

C_{gs} , C_{gd} , C_{ds} , R_{ds} , R_{gs} , and g_m are the classical GaAs pHEMT equivalent circuit elements.

g_{gd} (gate drain conductance), g_{gs} (gate source conductance) and R_{gd} (gate drain series resistance) are additional elements useful for a better accuracy under strong reverse gate-drain bias or forward gate-source bias.

Appendix C: Measurements of quality factors of one-port resonators using scattering parameters

Microwave resonators may be classified as either one-port or two-port circuits. Only the reflection coefficient is available for one-ports, so these are termed *reflection* resonators [119]. The use of s-parameter magnitudes is the easiest method by which to measure the bandwidth of a resonator, as these are the most familiar metrics used in the field. Although the classical 3-dB bandwidth method for determining the unloaded, loaded, and external Q is well established, one of its major disadvantages is that only three frequency points are used to determine the fractional bandwidth, even though the available frequency sweep may be composed of hundreds of data points. To remedy this problem, it was shown that *any* measured bandwidth can be related to the quality factor, not just the classical half-power, or 3-dB bandwidth [73].

The resonant microwave and mm-wave circuits exhibit sharp impedance variations at distinct frequencies. The behavior of the resonator may be described by an equivalent lumped resonant circuit in the vicinity of each resonance. The three basic parameters of each equivalent resonant circuit are: the unloaded resonant frequency f_o , the unloaded Q factor Q_o , and the coupling factor κ .

The input impedance as a function of frequency of a typical resonator describes a nearly perfect circle. Ginzton [108] describes in detail the analysis of measured circles on the Smith chart in order to deduce the values of Q_o and κ . The procedure described by Ginzton was developed at the time when impedances were measured with the use of

slotted lines. It was then appropriate to define “detuned short” and “detuned open” reference points on the slotted lines, and to interpret the data with respect to one or the other of these two points. The implied assumption was that the reactance of the external circuit can be neglected.

Nowadays, the impedance is measured much faster with the network analyzer [72]. The reference plane of such a measurement coincides with the end of transmission line leading to the resonator. The procedure to be followed here uses the data directly, such as shown on the network analyzer. Furthermore, it does not require that the reactance of the external circuit to be negligible.

The measurement is performed in the narrow range of frequencies around f_o , the resonant frequency under observation. So, for one port resonator the measurement using the reflection coefficient technique requires the following sequence [73] ,[90]:

- Measure the return loss at resonant frequency, the minimum S_{11} value, Γ_0
- Determine the coupling condition (under- or over-coupled), using $\text{Arg}(\Gamma_0)$. These two conditions are easily recognised on the Smith Chart: if the resonance appears to encircle the centre of the Smith Chart, it is overcoupled, otherwise it is undercoupled [120].
- Calculate the coupling factor κ from reflection coefficient as follows:

If under coupled,

$$\kappa_{under} = \frac{1 - |\Gamma_0|}{1 + |\Gamma_0|} \quad (C.1)$$

If over coupled,

$$\kappa_{over} = \frac{1 + |\Gamma_0|}{1 - |\Gamma_0|} \quad (C.2)$$

Calculate the return loss level in which the bandwidth Δf could be measured. It could be the 3 dB level or any level in general function of the power ratio p ($p=0.5$ for 3dB level, $p=0.1$ if the resonator is absorbing 10% from the incident power);

$$|\Gamma_{Unloaded}|^2 = \frac{(1-p)(1 \pm |\Gamma_0|)^2 + 4p\Gamma_0^2}{(1-p)(1 \pm |\Gamma_0|)^2 + 4p} \quad (C.3)$$

Go back to the measured data and determine the band width Δf at the calculated return loss level $|\Gamma_{unloaded}|$;

Extract the unloaded Q_o from its basic definition:

$$Q_o = \frac{f_o}{\Delta f} \quad (C.4)$$

Where These equations reduce to those given by [71] at half-power, $p = 0.5$.

Appendix D: Characteristics of the probes and cables employed in the measurements

D.1. Characteristics of the V-type cable used to measure the ribbon-bonded oscillators

D.2. Characteristics of the V-type cable used to measure the flip-chip mounted oscillators

W. L. GORE & ASSOCIATES, INC.

v2.10

SN 01454332

PN: 4F0BD0BC0480
17/May/2002 08:19 AM
Operator: RDK EA501
INS

Start = 0.125 GHz
Stop = 62.5 GHz

Maximum = -20.44 dB (1.21 : 1 SWR)

Maximum = -17.95 dB (1.29 : 1 SWR)

Maximum = -7.30 dB

Maximum = -7.25 dB

D.3. Characteristics of the GSG probe employed in the measurements.

Model 40A Performance Data ¹	
G,S,G Configuration	
Frequency Range:	DC to 40 GHz
Insertion Loss:	Less than 1.0 db to 40 GHz (.70 db typical)
Return Loss:	Less than 30 db to 4 GHz (35 db typical) Less than 20 db to 26 GHz (23 db typical) Less than 18 db to 40 GHz (20 db typical)
Crosstalk ² :	Less than 38 db to 40 GHz

¹ Specifications are for the P-style Model 40A Picoprobes with G,S,G configurations and pitches of between 50 and 300 microns. The C and T style Model 40A Picoprobes (see following page for mounting styles) have the same specifications except for insertion loss, which is less than 1.2 db (0.9 db typical).

² Crosstalk is measured using two probes contacting a bare sapphire substrate 100 microns apart.

Model 40A Performance Data ³	
G,S and S,G Configuration	
Frequency Range:	DC to 40 GHz
Insertion Loss:	Less than 2.0 db to 40 GHz (1.6 db typical)
Return Loss:	Less than 30 db to 4 GHz Less than 15 db to 26 GHz Less than 12 db to 40 GHz

³ Specifications are for the P-style Model 40A Picoprobes

S₁₁ Z
REF 1.0 Units
200.0mUnits/

START 0.045000000 GHz
STOP 40.000000000 GHz

Typical uncalibrated performance of a Model 40A-GSG-150-P Picoprobe. The top trace is the round trip return loss into a short which is twice the probe's insertion loss. The bottom trace is the return loss into a 50 ohm load.

Crosstalk performance of two Model 40A-GSG-150-P Picoprobes while contacting a bare sapphire substrate with spacings of 100, 200, and 400 Microns.

LEFT: A Smith Chart showing the calibrated response of a Model 40A-GSG-150-P while contacting a coplanar offset short. The LRM method was used for calibration.