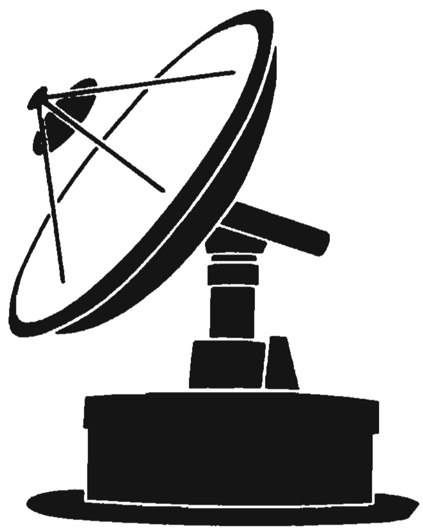
OSCILLATOR DESIGN USING MODERN NONLINEAR CAE TECHNIQUES

Network Measurements Division
1400 Fountaingrove Parkway
Santa Rosa, CA 95401-1799

AUTHORS:
Michal Odyniec
Jeff Patterson
Robert D. Albin
Andy Howard



RF & Microwave
Measurement
Symposium
and
Exhibition

HEWLETT
PACKARD

ABSTRACT

An overview of the design of RF and microwave large signal oscillators highlights the use of modern nonlinear CAE tools.

Traditional forms of oscillator analysis using S-parameters are first reviewed. Then the techniques using the new harmonic balance nonlinear simulator are outlined. An RF VCO and a microwave YIG oscillator are used as case studies. In both cases design strategies are presented along with simulated data compared to actual measurements.

The authors thank Ms. Mary Fewel, Mr. Roger Swearingen, and Mr. Jim Fitzpatrick for helpful comments made on the manuscript.

AUTHORS

Michal Odyniec received his MS in Applied Mathematics and PhD in Electrical Engineering, both from the Technical University of Warsaw, Poland. From 1981-1984, Michal was a visiting faculty member at the University of California at Berkeley. In 1985 he joined Microsource, Inc. where he became a project leader responsible for designing wide-band microwave oscillators. In 1989, Michal joined HP's Network Measurements Division in Santa Rosa, California where he works on nonlinear CAE.

Jeff Patterson joined HP in 1981 after receiving his BSEE from Rice University in Houston, Texas. He worked as a production engineer from 1981-1985 with responsibility for various families of spectrum analyzers. From 1986 to 1988, Jeff was a member of the new product introduction team for the portable family of spectrum analyzers. Since then, he has worked in R&D on low noise oscillator design.

Robert D. (Dale) Albin received his BSEE degree in 1977 from the University of Texas at Arlington, and MSEE degree from Stanford University in 1980. After joining the Microwave Technology Division of HP in 1977, Dale was a Production Engineer working on GaAs FET testing and processing. He then joined HP's Network Measurements Division where he was a project leader on millimeter source modules and project manager for lightwave sources and receivers. He is currently working on YIG oscillators.

Andy Howard is an R&D engineer at HP's Network Measurements Division in Santa Rosa, California. Andy earned his BSEE in 1983 and MSEE in 1985, both from the University of California in Berkeley. While earning his MS, Andy learned Japanese and spent nine months working at NEC's Central Research Laboratories in Japan where he investigated an analog-to-digital conversion algorithm. After joining HP in 1985, Andy developed scalar network detectors and a bridge. He is currently working on YIG oscillator development.

Slide 1 INTRODUCTION

OSCILLATOR DESIGN USING NON-LINEAR CAE

This presentation introduces basic concepts of oscillator analysis and design. It also shows new applications of nonlinear CAE to these circuits.

Slide 2 CONTENTS

CONTENTS

- Review of RF and Microwave Oscillators
- Methods of Oscillator Analysis

The paper consists of two parts:
1. In the first part, we review transistor and diode oscillators.
2. In the second part, the main body of the paper, we present the fundamental linear and nonlinear methods of analysis and design. The results are illustrated by computer analysis of a simple circuit and also of real oscillators. Different methods of design are shown and compared with measured data.

Slide 2a

CONTENTS

- ➡ Review of RF and Microwave Oscillators
- Methods of Oscillator Analysis

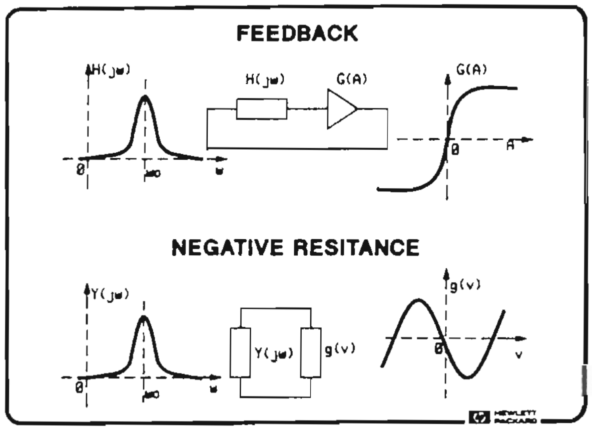
Slide 3 REVIEW OF OSCILLATORS

REVIEW OF OSCILLATORS

- RF: Armstrong, Colpitts, Hartley, Crystal
- Microwave: SAW, DRO, VCO, YIG
- mm-Wave: Gunn, Impatt

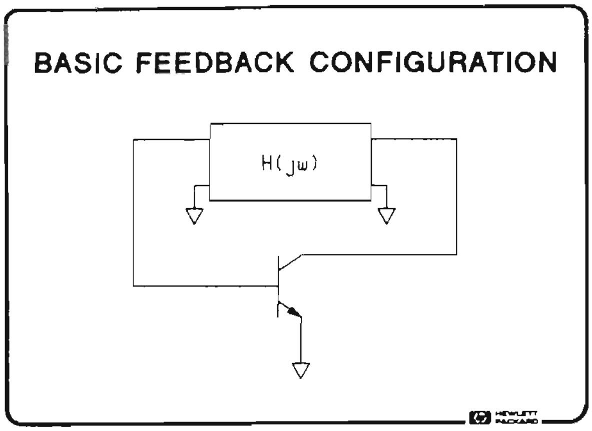
We present basic oscillator structures and review practical implementations for frequency bands from RF to mm-waves.

Slide 4 FEEDBACK VS. NEGATIVE RESISTANCE



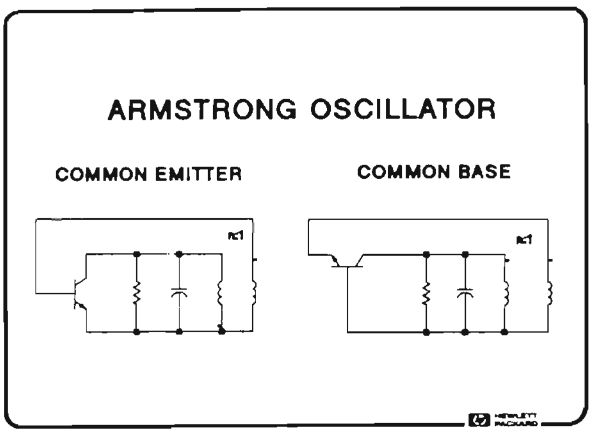
There are two basic approaches to oscillator design:
1. The first views an oscillator as a feedback system with a nonlinear amplifier and a selective filter.
2. The other divides an oscillator into an active circuit and a resonator (tank circuit). Most oscillators can be analyzed in either way; convenience dictates which approach to choose.

Slide 5 BASIC FEEDBACK STRUCTURE



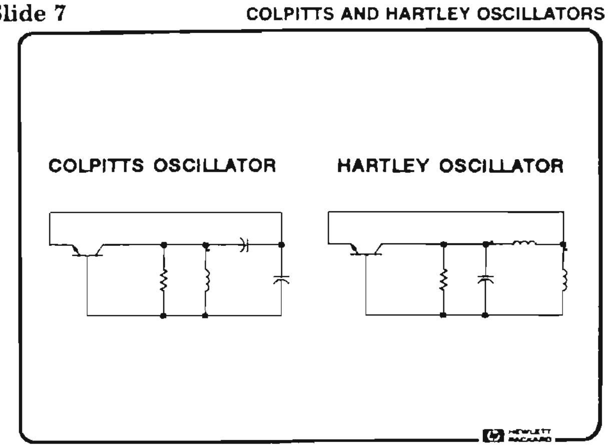
Let us start with the feedback description which allows for the uniform treatment of many transistor oscillators. We shall review transistor oscillators starting from the classic (but still used) structures of Armstrong, Colpitts and Hartley. Then we shall see how different resonator types extend the oscillation frequency Finally, we review wide-band tunable oscillators.

Slide 6 ARMSTRONG OSCILLATOR



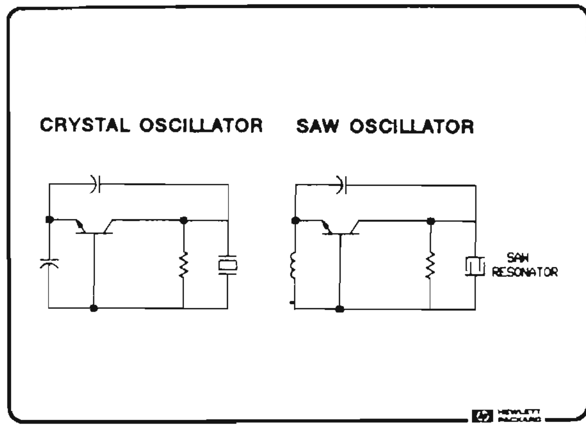
In a classic Armstrong oscillator (in both common-emitter and common-base configurations) one can easily determine the feedback path, the nonlinear amplifier, and the frequency-determining LC filter.

Slide 7 COLPITTS AND HARTLEY OSCILLATORS



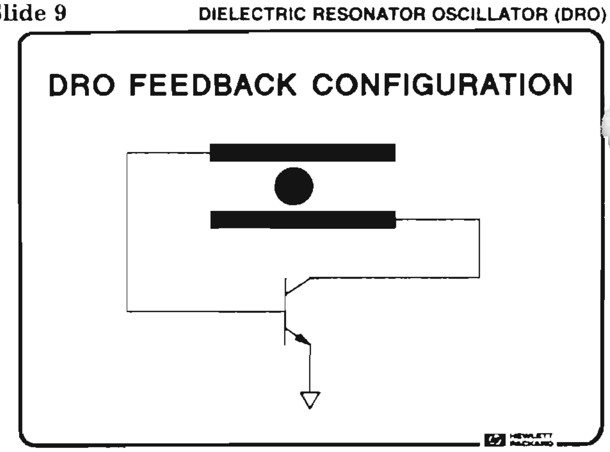
The Colpitts replaces the coupled inductors with a capacitive RF transformer. The Hartley is dual to the Colpitts. Interestingly, those basic structures were introduced just after the invention of the active device (the triode, which was called "audion" at that time) and are still used [1].

Slide 8 CRYSTAL AND SAW OSCILLATORS



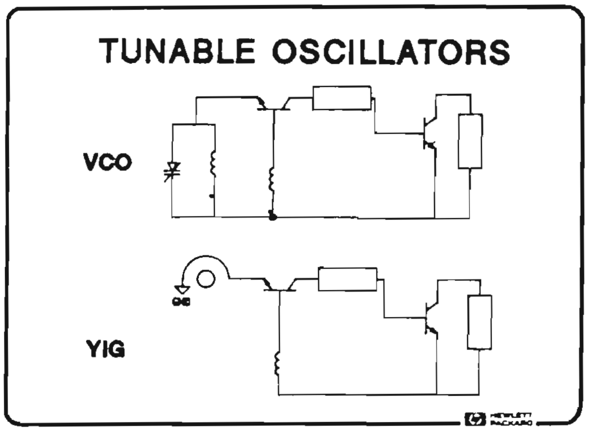
Introducing a quartz crystal into the feedback loop provides a very stable oscillator. High frequency stability of the oscillators makes them attractive for use not only at RF but also at microwave frequencies using frequency multipliers. Crystal oscillators are usually limited to 100 MHz. For the frequency range to 2 GHz, SAW (surface acoustic wave) resonators are used. Dielectric and cavity resonators are used above 2 GHz. The SAW oscillator uses a SAW delay line which acts as a selective filter in the feedback path.

Slide 9 DIELECTRIC RESONATOR OSCILLATOR (DRO)



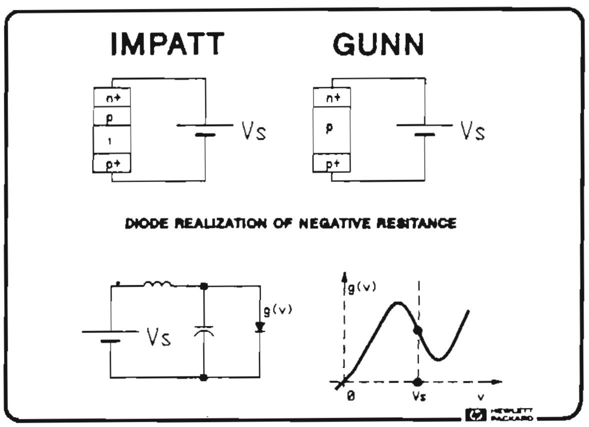
In dielectric resonator oscillators, the selective feedback filter is realized by a dielectric resonator coupled to a microstrip or a stripline [2,3]. The dielectric resonator has high Q (unloaded Q=7000 was reported at 6 GHz [2]), resulting in high frequency stability, and it resonates in the range of 1 to 60 GHz. Frequency depends on resonator dimensions, resonator position, and substrate and air gap thickness. Thus, frequency tuning can be accomplished by mechanically changing the gap's dimensions.

Slide 10 TUNABLE OSCILLATORS VCO AND YIG



In many applications (test equipment, electronic warfare, and to a lesser extent in communications), it is important to tune oscillators over a wide frequency range. This is achieved by varactor diodes and YIG resonators [2,3]. In RF and microwave ranges, a varactor diode serves as a voltage-controlled capacitor providing tuning capabilities. It has extremely fast tuning speeds. Its Q, however, is low ($Q < 50$), and this results in poor frequency stability. The varactor is typically used with LC elements for wide (1.5 octave) tuning. It can also be used with a dielectric resonator for narrow tuning with better Q. The other way to provide very wide tuning at microwave frequencies, and also good frequency stability, is to use YIG resonators. A YIG sphere, when installed in uniform magnetic field, behaves like a resonator with 1000-8000 unloaded Q. When the field is varied, the resonant frequency changes (with excellent linearity, approximately 0.1% per octave). YIG resonators are useful in the 1-60 GHz frequency range (limited by magnetic field saturation). Varactors are useful in the range from RF to 30 GHz.

Slide 11 GUNN AND IMPATT OSCILLATORS



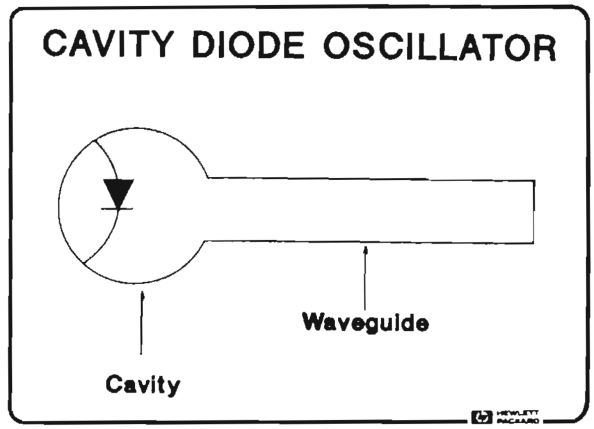
Transistors have both low noise and high efficiency but their practical frequency range is limited to 40-50 GHz. Therefore, for higher (mm-wave) frequencies, Gunn and IMPATT diodes are commonly used [3,4]. At such high frequencies, amplification is a serious problem, and power efficiency becomes an important oscillator characteristic. Gunn diode oscillators, when compared to IMPATT, have high frequency stability but poor power efficiency. The Gunn diode oscillator can operate in either the so called "Gunn mode", when it produces pulses, or in the LSA mode (limited space- charge accumulation) when it presents a negative resistance without producing pulses. There are commercially-available, mechanically-tunable Gunn oscillators that produce: .5-1.0 W in the X-band (8-12GHz) with 10% efficiency (GaAs) .2 W at 40 GHz with 2% efficiency (GaAs) .2 W at 66 GHz with 8% efficiency (InP) .1 W at 100 GHz with 2% efficiency (InP)

An IMPATT diode (or a diode in IMPATT mode) oscillates to 100 GHz, and even to 400 GHz (using higher harmonics).

The output powers are about: 10 W at 10 GHz with 20% efficiency (GaAs or Si) 1.5 W at 50 GHz with 10% efficiency (GaAs or Si) 60 mW at 100 GHz with 1% efficiency (GaAs) 50 mW at 220 GHz with 1% efficiency (Si)

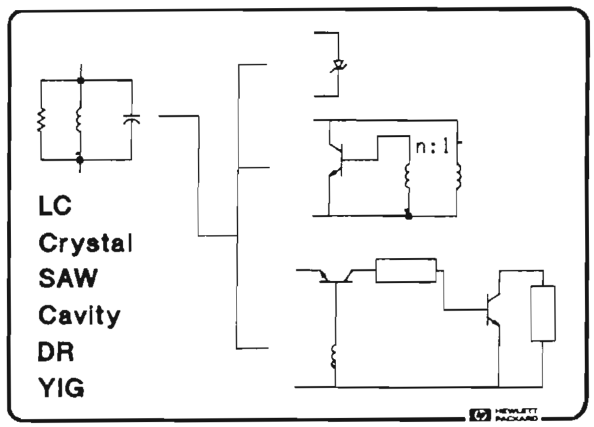
Both Gunn and IMPATT diodes can be viewed as a negative resistance in parallel with capacitance.

Slide 12 CAVITY OSCILLATORS



Cavity resonators are often used with diodes, provide high Q, and consequently, high frequency stability; but, because of resonator dimensions, they are only practical for frequencies above 4GHz.

Slide 13 NEGATIVE RESISTANCE CONFIGURATION



Note that each of the oscillators we reviewed above can be split into a resonator and a "negative resistance". One can argue that indeed the oscillator design consists of providing an appropriate resonator and a negative resistance (or an amplifier and a feedback filter). Therefore, depending on application, we choose between LC, crystal, cavity, SAW, dielectric, or YIG resonators,(we can also add a varactor for tuning). We also choose between diodes (tunnel, Gunn, or IMPATT) or transistors (BJTs, FETs,or HEMTs) to provide the negative resistance.

Slide 14

CONTENTS

- Review of RF and Microwave Oscillators
- ➡ Methods of Oscillator Analysis

Slide 15 NONLINEAR METHODS OF OSCILLATOR DESIGN

METHODS OF OSCILLATOR ANALYSIS

- Review of Non-linear Methods
- Small and Large Signal Feedback
- Small Signal S-parameters
  - Design Examples
- Large Signal S-parameters
  - Design Examples
- Oscport Analysis
  - Design Examples

As we have seen above, oscillators can be modeled in both feedback and negative resistance configurations. For each, we will review the methods of nonlinear analysis and design.

Slide 15a

METHODS OF OSCILLATOR ANALYSIS

- ➡ Review of Non-linear Methods
- Small and Large Signal Feedback
- Small Signal S-parameters
  - Design Examples
- Large Signal S-parameters
  - Design Examples
- Oscport Analysis
  - Design Examples

Slide 16 NONLINEAR METHODS - REVIEW

NON-LINEAR METHODS

- Direct Time-Domain Simulation
- Approximate Methods
  - "Limited" Signals
  - Periodic Signals

The nonlinear differential equations cannot (except in ver special cases) be solved exactly. Thus nonlinear analysis use either direct circuit simulation or the approximate methods The former have limited use for microwave circuits whe these include distributed elements. Therefore we shall con centrate on the approximate methods.

Slide 17 APPROXIMATE METHODS - LIMITED SIGNALS

"LIMITED" SIGNALS

Power / Voterra Series

$$g(V_0 + v) = g(V_0) + g'(V_0)\,v + \frac{g'(V_0)^2}{2!} + \cdots$$

Local Linerarization

$$g(V_0 + v) = g(V_0) + g'(V_0)\,v$$

Slide 18 APPROXIMATE METHODS - PERIODIC SIGNALS

PERIODIC SIGNALS

Harmonic balance

$$x(t) = \sum_{n=1}^{N} X_n \exp(jwnt)$$

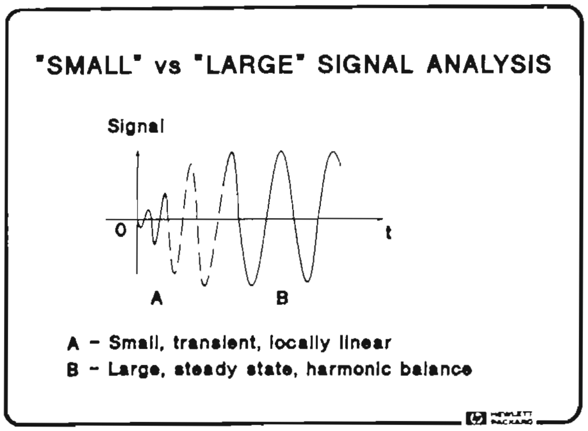
$$g(x(t)) = \sum_{n=1}^{N} G_n(X_0, X_1, X_2, \ldots, X_N)\exp(jwnt)$$

Averaging

$$x(t) = A(t)\cos(wt + phi(t))$$

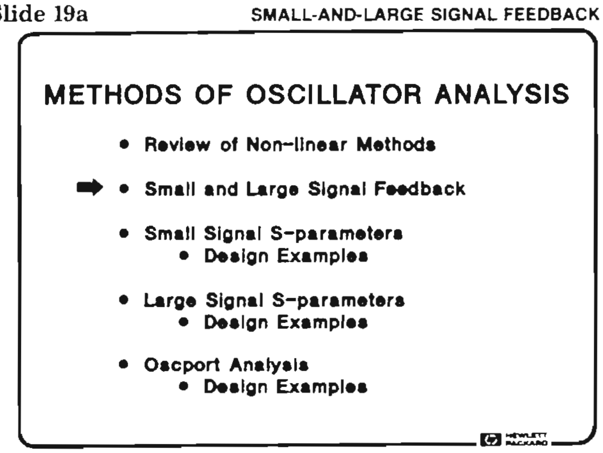
The approximate methods fall into two classes: 1. We assum that the considered signals are "limited" so that the nonline ar characteristics can be expanded into Volterra series. In practice, we consider polynomial, rather than series, expan sion with a finite (preferably low) number of the Vo components. 2. The second approach consists of conside steady-state periodic waveforms. In the harmonic balanc method, we look for solutions that are represented by finit Fourier expansion; in the averaging method, we look for sinusoids with "slowly" varying amplitudes and phases. Below we shall concentrate on the method of LOCAL LINEARIZATION (power expansion which ends at the linear term) and that of HARMONIC BALANCE.

Slide 19 SMALL-AND-LARGE-SIGNAL ANALYSIS

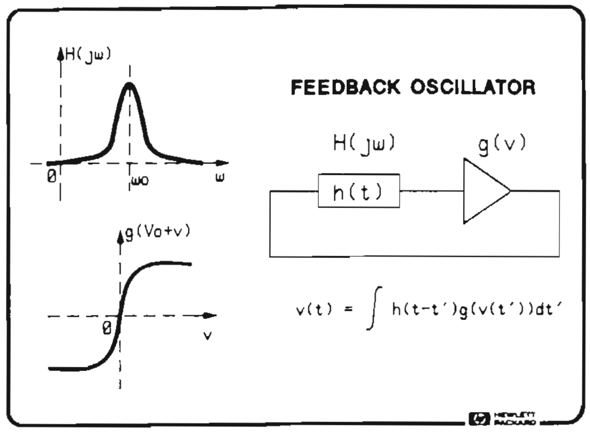


The simplest and most widely used analysis method consists of LOCAL LINEARIZATION, i.e. of considering only the first coefficient in Taylor expansion. Thus the system becomes locally linear about the operating point. The method is typically called "SMALL SIGNAL" analysis. The other popular method is that of HARMONIC BALANCE, in which the signal is replaced by a finite number of its harmonics. Since general systems can be easily presented in the feedback description, we shall use the latter to present the two methods. Then we analyze specific examples in the negative resistance setting.

Slide 19a SMALL-AND-LARGE SIGNAL FEEDBACK



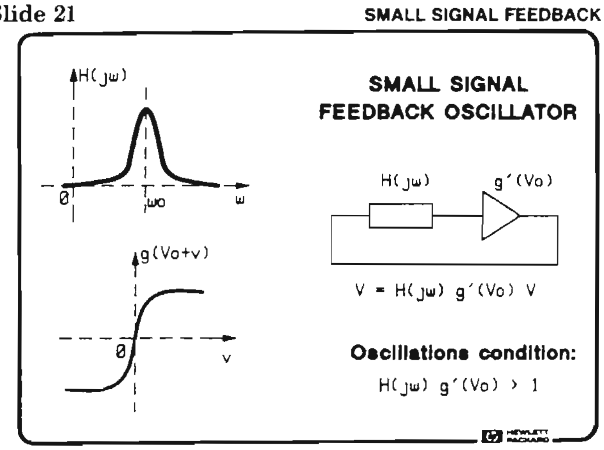
Slide 20 FEEDBACK DESCRIPTION



Consider a feedback representation of an oscillator circuit, which consists of a linear "filter" (specified by the impulse response:h(t) and the transfer function H(s)) and a nonlinear, memoryless element(s): g(Vo+v)).
In time domain, the circuit is represented by a convolution: $v(t) = [h(t)*g(Vo+v(t))]$
In frequency domain, it is described by:
1. $Vo = H(0)g(Vo)$ for dc analysis
2. $V = H(s)g'(Vo)V$ for "small-signal", locally-linear analysis
3. $Vn = H(jnw)Gn(Vo,V1,V2,...,VN)$ $n = 1,2,...,N$ for harmonic-balance analysis, where Vn and Gn(.) are the Fourier coefficients of v(t) and g(v(t)).

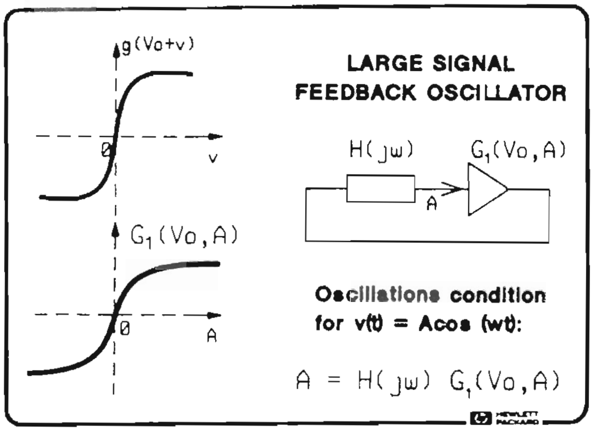
The first two equations are obtained from power series expansion; the last one from harmonic expansion. Indeed, substituting $g(Vo+v) = g(Vo) + g'(Vo)v$ into the convolution, we get: $Vo+v = h(t)*g(Vo) + h(t)*[g'(Vo)v] = H(0)g(Vo) + g'(Vo)[h(t)*v]$
The first term of which yields equation 1.; the second, equation 2. Similarly, by expanding v(t) and g(Vo+v(t)) into Fourier series, we get $Vo+v(t) = [h(t)*(Go + G1exp(jwt) + G2exp(j2wt) + ...)] = = [H(0)Go + H(jw)G1exp(jwt) + H(j2w) G2exp(j2wt) + ...)]$ which yields the equation 3.

Slide 21 SMALL SIGNAL FEEDBACK



In the small signal analysis we assume that all waveforms consist of "constant component + small time-varying" signal, and we expand the nonlinear characteristic into Taylor series about the operating point Vo: $g(Vo+v) = g(Vo) + g'(Vo)v + g''(Vo)v'2/2! + ...$ Since the time-varying signal is "small", we can neglect the second and higher order terms in Taylor expansion. Consequently we get a linear feedback system. The system is unstable (and consequently oscillates) when the open loop gain is larger than one: $H(jw)g'(Vo) > 1$
This condition can also be obtained from Nyquist stability criterion.

Slide 22 LARGE-SIGNAL FEEDBACK



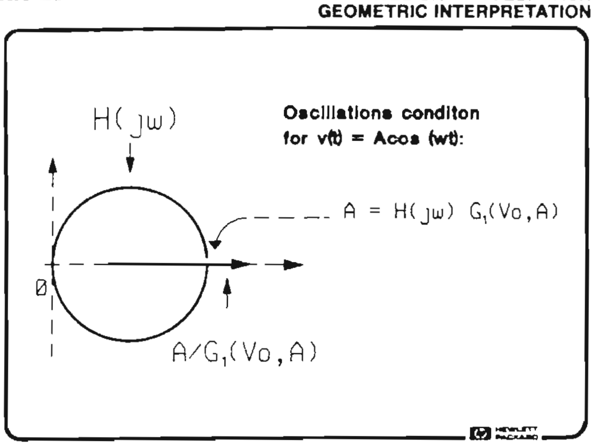
In the harmonic-balance method, we look for steady-state periodic oscillations. When the oscillations are represented by finite Fourier series, the circuit equations read: Vn = H(jnw)Gn(Vo,V1,V2,...,VN) n = 1,2,...,N (23)

1. Special case: v(t) ~ A cos(wt)
When the oscillations are close to sinusoidal (which takes place in most oscillators), we can neglect all harmonics except the first one. In this case, the harmonic- balance equations simplify to a single (complex) equation with unknown frequency and amplitude (we choose the phase so that the amplitude is real).

A = H(jw)G1(Vo,A) (23a)

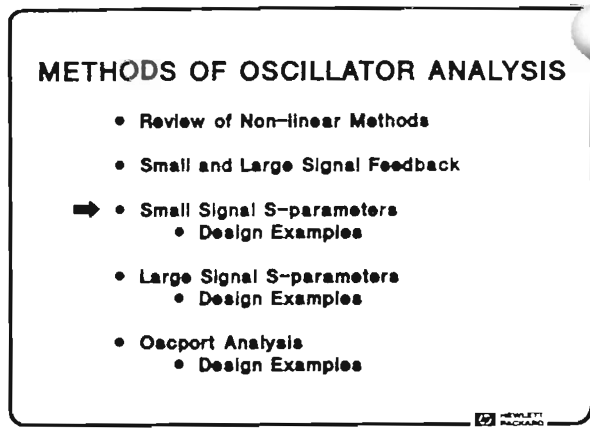
The solution of (23a) has simple physical interpretation: the large signal gain around the feedback loop equals to one: 1 = H(jw)[G1(Vo,A)/A].

Slide 23 LARGE-SIGNAL FEEDBACK GEOMETRIC INTERPRETATION

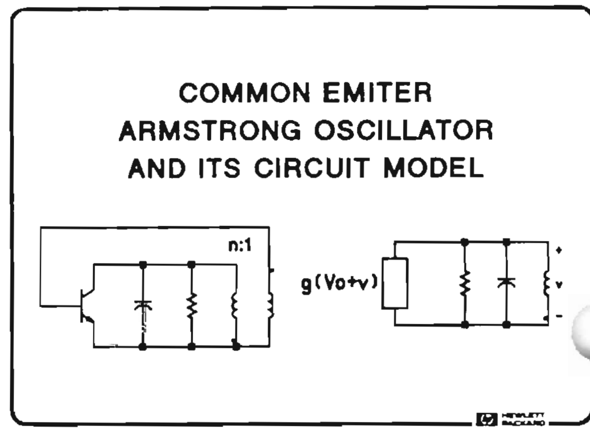


Geometrically the solution of (23) can be found at the intersection of two curves: one depending on frequency - H(jw), the other on amplitude - A/G1(Vo,A).

Slide 23a

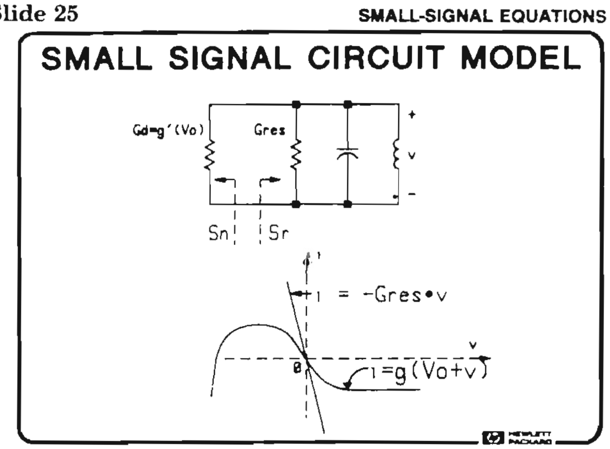


Slide 24 ARMSTRONG'S OSCILLATOR CIRCUIT MODEL



Let us review small-signal-analysis in the "negative resistance" setting. We start with the Armstrong oscillator, which can be easily reduced to a simple RLC circuit with nonlinear resistance. We shall later use the same oscillator for large signal analysis.

Slide 25 SMALL-SIGNAL EQUATIONS



As we already explained, when the time-varying signals are "small", we can expand the nonlinear characteristics into Taylor series neglecting the second and higher order terms of the expansion:

g(Vo+v) = g(Vo) + g'(Vo)v

Consequently we obtain circuit equations which are linear in "v":

dc:

0 = V - Vo

0 = Io - g(Vo)

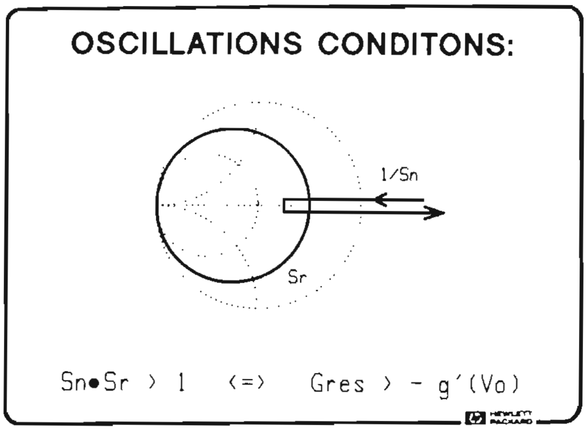
ac:

Ldi/dt = - v

Cdv/dt = i - g'(Vo)v - Gres v

Clearly the circuit oscillates when its total conductance (g'(Vo)+Gres) is negative.

Slide 26 S-PARAMETER DESIGN

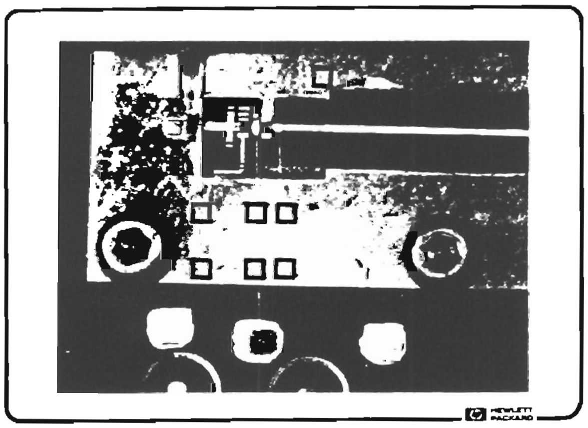


Microwave circuits are easy to analyze in terms of S-parameters, which are shown here for the small-signal oscillator with varied bias.

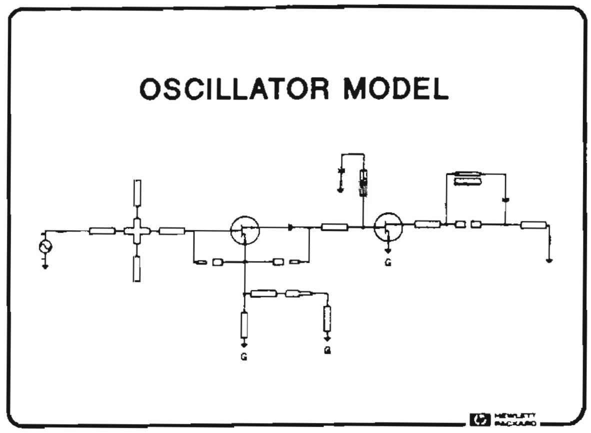
The oscillator design is based on the small-signal relationship: Sn*Sr >1, which says that if the double-reflected signal is bigger than the original one, then the circuit oscillates. Note that this condition is equivalent to the conditions: Gd(Vo)+Gres < 0 (total resistance negative) obtained in ac analysis, and g'(Vo)H(jw) >1 obtained in feedback analysis above.

In oscillator design the aim is to find bias for which Sn*Sr > 1 holds over the desired frequency band.

Slide 27 8-20 GHZ YIG OSCILLATOR

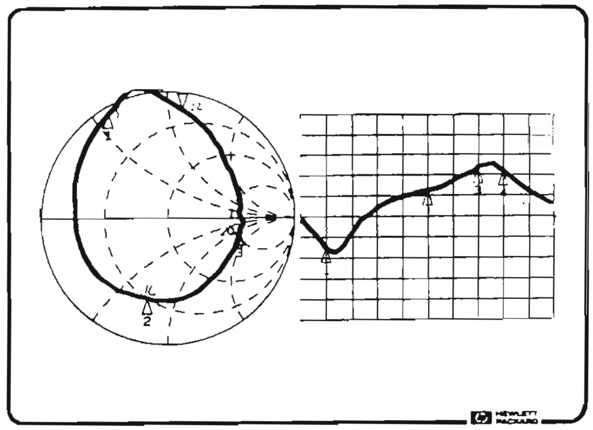


Slide 28 8-20 GHZ OSCILLATOR - DIAGRAM

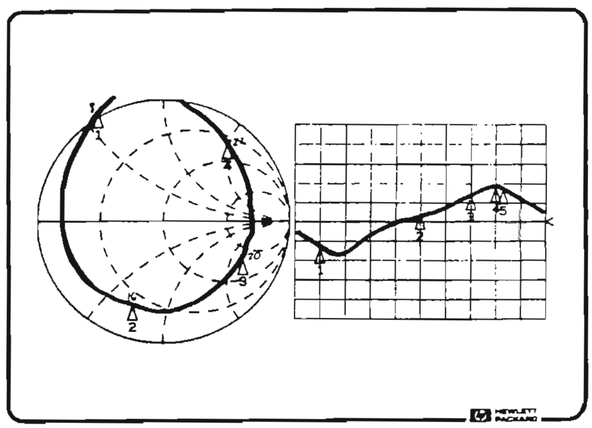


As a second example, let us consider the bias dependent small-signal design of an 8-20 GHz YIG oscillator.

Slide 29 SMALL-SIGNAL S-PARAMETER DESIGN - SIMULATION



Slide 30 SMALL-SIGNAL S-PARAMETER DESIGN



The oscillator design is based on the small-signal relationship: Sn*Sr >1 and Sn dependence on bias. Since the YIG resonator has high Q, the Sr has the magnitude close to 1 and a fast varying phase. Consequently, the oscillations exist for those frequencies for which 1/Sn<1. The designer's aim is to choose FETs and bias conditions so that, for the desired frequency range, 1/Sn fits inside the Smith chart, (rigorously speaking, it fits inside the resonator characteristic which is close to the unit circle). The measured small-signal S-parameters closely follow the simulated ones.

Slide 30a

METHODS OF OSCILLATOR ANALYSIS

- Review of Non-linear Methods
- Small and Large Signal Feedback
- Small Signal S-parameters
  - Design Examples
- ➡ Large Signal S-parameters
  - Design Examples
- Oscport Analysis
  - Design Examples

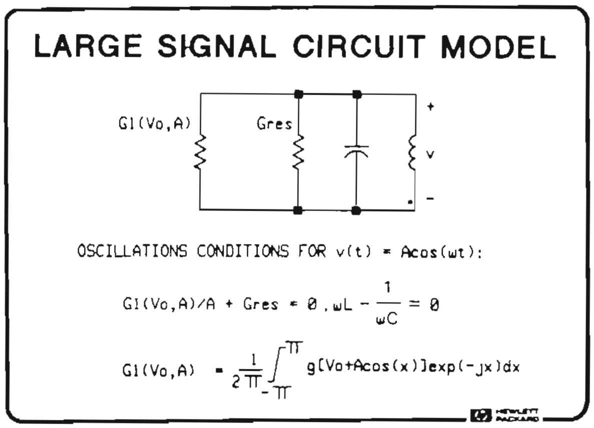
When discussing feedback systems we found that the harmonic balance circuit equations have the form:

Vn = H(jnw)Gn(Vo,V1,V2,...,VN) n = 1,2,...,N (23)

In oscillator analysis the frequency is not known a priori. This makes the numerical analysis of the equation (23) particularly difficult.

We present below two ways of overcoming this difficulty. One, which generalizes the geometrical approach presented above for feedback systems, consists of analysis of large signal S-parameters. The other introduces the "oscport" device, which allows direct simulation of oscillator circuits.

Slide 31 HARMONIC BALANCE ANALYSIS



The harmonic-balance equations for the Armstrong's oscillator take the form:

jnwLIn = - Vn

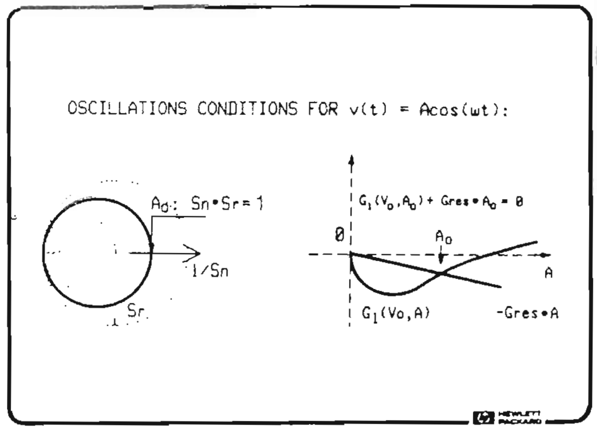
GresVn + jnwCVn = In - Gn(Vo,V1,...,Vn,...)

If we neglect all harmonic coefficients except the first (we can do so because YIG is a high Q resonator), then the equations reduce to:

(Gres + jwC + 1/jwL)V1 = - G1(Vo,V1)

where Vo is a parameter, w and V1 are unknown real numbers (we choose oscillations phase so that V1 is real). For the LC oscillator, G1 is real valued, and we easily obtain oscillations with frequency wo = 1/sqrt(LC) and amplitude V1 = Ao, where G1(Vo,Ao) = -Gres.

Slide 32 LARGE-SIGNAL S-PARAMETERS FOR THE LC OSCILLATOR



Since signals in nonlinear circuits are sums of sinusoids, the concept of impedance and that of S-parameters are not obvious. A natural way to define the "large-signal" impedance, or S- parameters, would consist of considering the first Fourier coefficients (the fundamentals) of all waveforms, and defining with them incident and reflected waves. Suppose that the voltages and currents have the form:

v(t) = Vo + (V1)cos(wt+p1) + (V2)cos(2wt+p2) + ...

i(t) = Io + (I1)cos(wt+q1) + (I2)cos(2wt+q2) + ...

and let V1 = (V1)exp(jp1), I1 = (I1)exp(jq1).

We can now define the "large-signal" incident and reflected waves:

incident: a = (V1 + Zo*I1)/(2*sqrt(Zo))

reflected: b = (V1 - Zo*I1)/(2*sqrt(Zo))

After that, the definition of large-signal S-parameters follc naturally: Sik = bi/ak (with al = 0 for 'l' unequal to 'k

Similarly one defines the "large signal" impedance:

Z = V1exp(jp1)/I1exp(jq1)

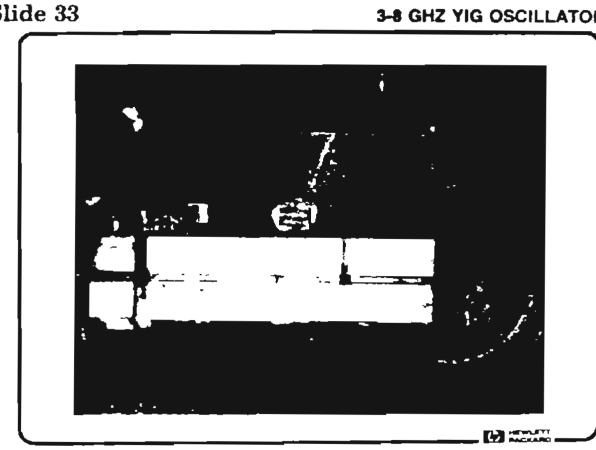
Note that:

1. The relationship S11 = (Z-Zo)/(Z+Zo) holds for the "large-signal" definition.

2. The large signal S-parameters vary with amplitude and the small signal ones equal to the limit value obtained for amplitude converging to zero.

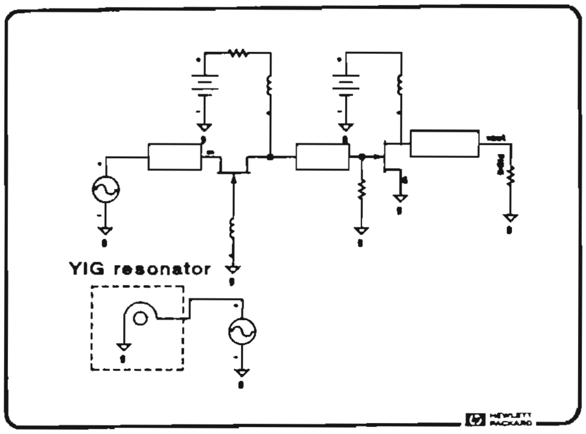
3. The large signal parameters are less dependent on bias The large signal S-parameters defined as above can be used for steady- state oscillator design. This is based on the large signal relationship: Sn*Sr = 1 The intersection point of 1/Sn and Sr gives us amplitude and frequency of actual oscillations; we can also evaluate the phase noise from it .

Slide 33 3-8 GHZ YIG OSCILLATOR

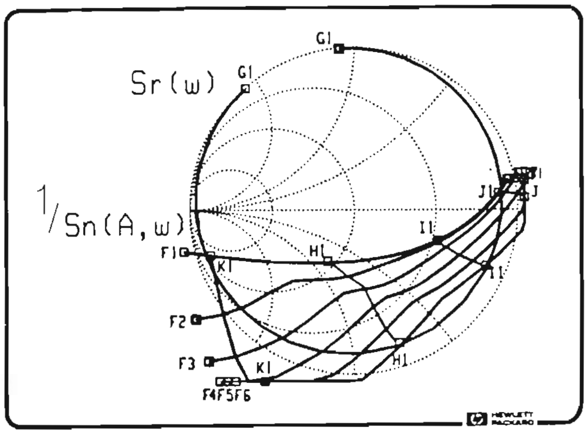


For design purposes, the circuit is split into the resonator part and the active part.

Slide 34 OSCILLATOR DIAGRAM



Slide 35 LARGE-SIGNAL S-PARAMETERS

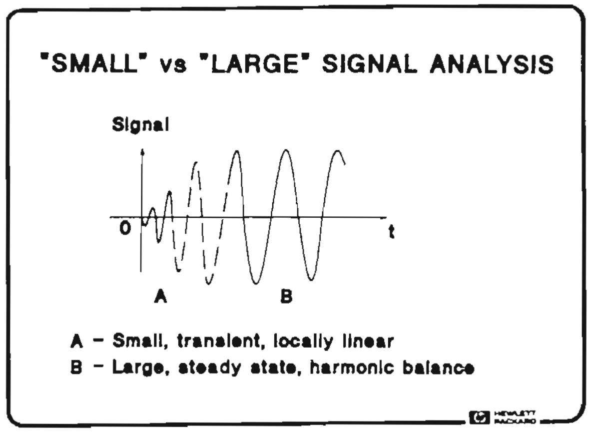


The multi-octave oscillator was designed based on the large-signal relationship: Sn*Sr = 1 The design aims were: 1. To select FETs and bias conditions for a wide band oscillator. 2. Estimate the power handled by the YIG sphere. 3. Estimate phase noise.

The circuit was split into the resonator part and the active part. For the active part, the harmonic-balance analysis provided a family of large-signal S-parameters swept in power and frequency. The curves F1-F6 show S-parameters for frequency swept from 2.5 to 9.0 GHz, with amplitudes fixed at:.2, .8, 1.4, 2.0, 2.6, 3.2.V (note that the small amplitude curve (F1) coincides with the small signal S- parameters). Similarly the curves K1,H1,I1,J1 show the S-parameters swept in amplitude with frequencies respectively fixed at 2.75, 4.25, 6.5, 8.25 GHz. The S-parameter characteristics are overlapped with the resonator plot (which can be obtained independently with the harmonic-balance as well as the ac analysis).

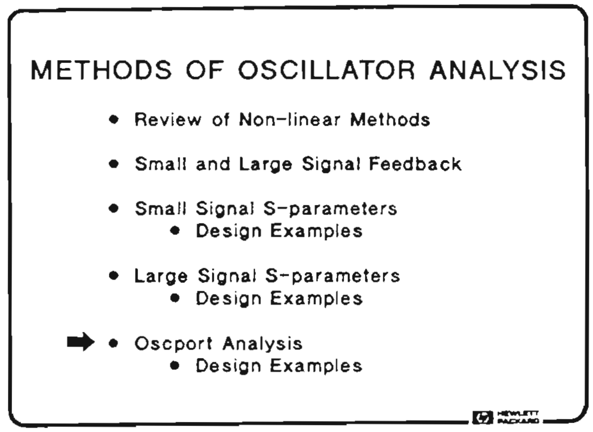
The intersection points of 1/Sn and Sr give us amplitude and frequency of oscillations; we conclude that the oscillations start just below 3.0 GHz and cease just above 8.2 GHz. At the low end, the voltage increases slowly with frequency, and at the high end, it changes fast. At frequencies between 4.25 and 6.5 GHz, the voltage is higher than 3.2 V. Once we know voltage and the circuit impedance, we can calculate power delivered to the YIG sphere (in our circuit,for example, the power varies from a few mW at 4.25GHz to tens of uW at 8.25 GHz). Consequently we can choose a sphere with appropriate power handling capability. (Let us note that, as the circuit impedances change with frequency, the power changes do not necessarily follow those of the voltage.) The phase noise is determined by the intersection angle between Sr and 1/Sn. Thus, we can estimate its level from the plots. For example, in the analyzed oscillator, phase noise is high at ~.0 GHz and diminishes with increasing frequency.

Slide 35a SMALL- AND LARGE-SIGNAL ANALYSIS

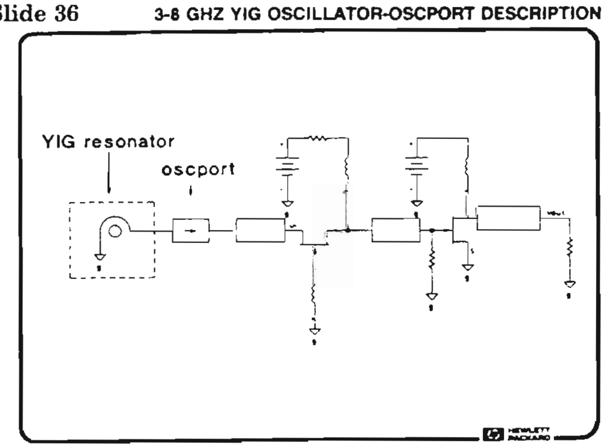


Note that although the formulae for small and large signal design are similar, they describe different physical phenomena.

Slide 35b OSCPORT ANALYSIS

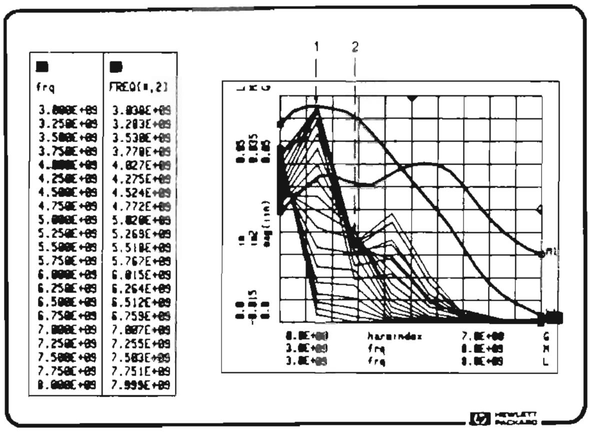


Slide 36 3-8 GHZ YIG OSCILLATOR-OSCPORT DESCRIPTION



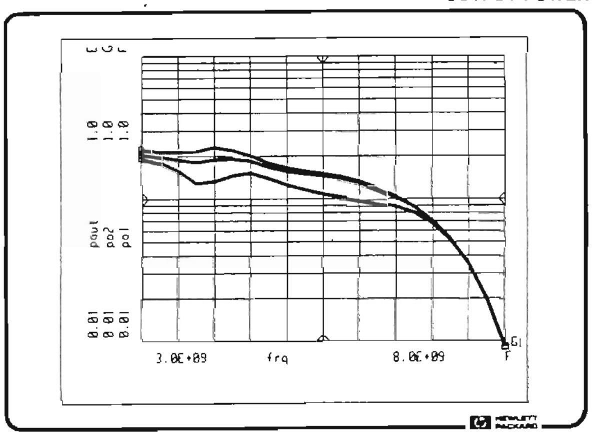
The third way of oscillator design consists of direct numerical analysis of circuit equations. In order to be able to calculate the unknown frequency the HP85150-provides a new device - the "oscport" probe; (it also has an analogous probe for small signal analysis - the "osctest", which we shall not discuss here). When inserted into the oscillator's feedback path, the oscport performs the harmonic balance analysis of the oscillator. Specifically, we can calculate the frequency, output power, or the power across the resonator (or at any point in the circuit).

Slide 37 3-8 GHZ YIG OSCILLATOR-FREQUENCY AND HARMONICS



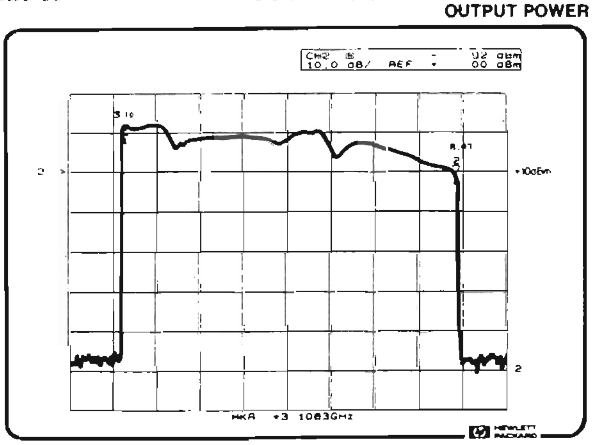
The simulation was performed with resonator frequency swept- from 3 to 8 GHz. Oscillations frequency voltage spectrum across the resonator, and amplitudes of its first and second harmonics are shown above for varying resonator frequency.

Slide 38 3-8 GHZ YIG OSCILLATOR-SIMULATED OUTPUT POWER



The output power was calculated for the fundamental, second and fifth harmonics.

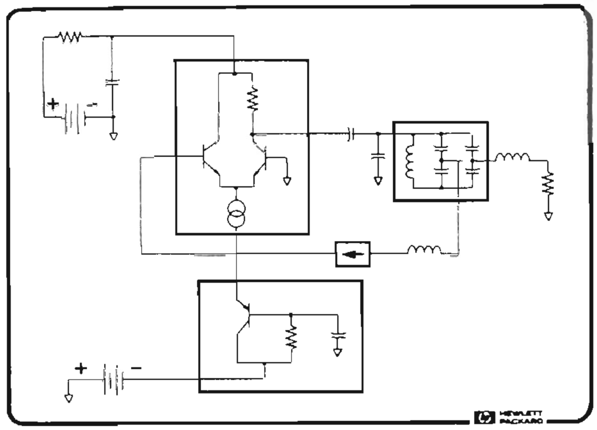
Slide 39 3-8 GHZ YIG OSCILLATOR-MEASURED OUTPUT POWER



As predicted by the large-signal S-parameter analysis, the oscillations start at 3 GHz and cease after 8 GHz. The output power (simulated with the oscport) reaches 20-23 dBm between 3-?? GHz and drops to 10 dBm at 8 GHz.

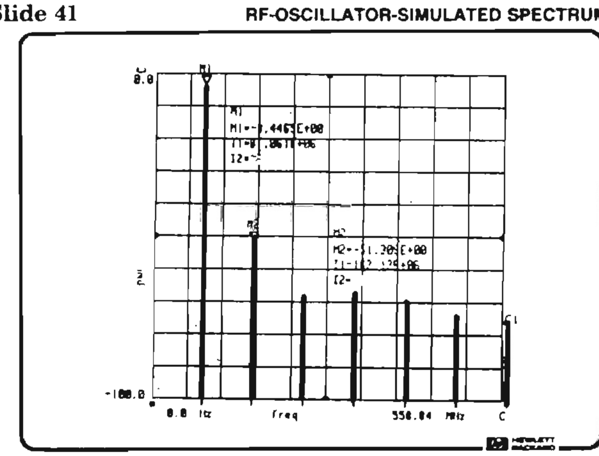
The YIG oscillator examples considered above were taken at early design stages when only global qualitative behavior, rather than exact agreement with measurements, was of interest. Consider now an oscillator which was simulated at final design stages.

Slide 40 RF OSCILLATOR-OSCPORT DESCRIPTION

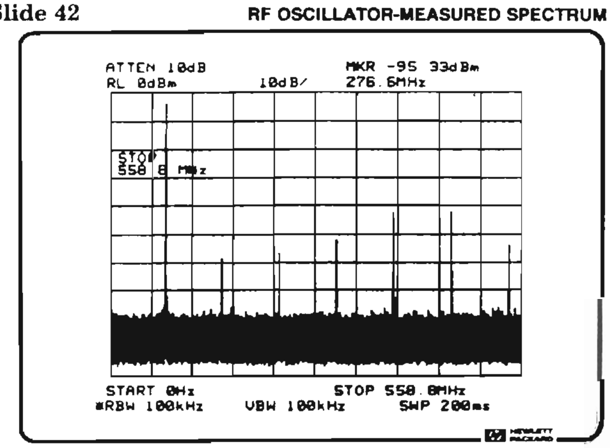


A varactor-tuned RF oscillator was built and simulated using the "oscport". Note that the oscillator diagram includes subcircuits of the differential pair and the tunable resonator.

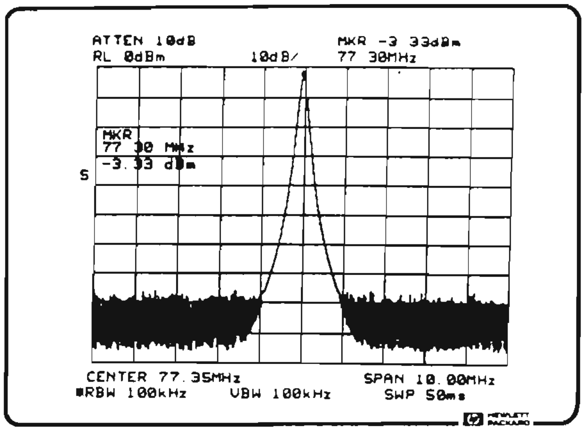
Slide 41 RF-OSCILLATOR-SIMULATED SPECTRUM



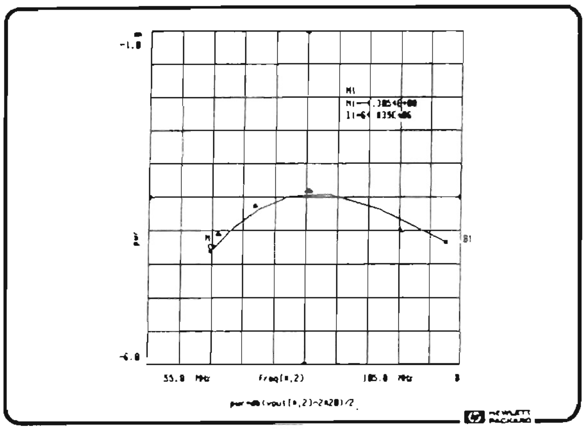
Slide 42 RF OSCILLATOR-MEASURED SPECTRUM



Slide 43 RF OSCILLATOR-MEASURED FREQUENCY



Slide 44 RF OSCILLATOR-OUTPUT POWER



Simulation results and the corresponding measurements of the actual circuit are shown: 1. Power of the fundamental agrees within .5dB over the band; 2. Second harmonic below the fundamental is 48 dB when simulated, 54 dB when measured; 3. Measured (77.3 MHz) and simulated (81.0 MHz) frequency differ by 5%. We attribute the discrepancy to running simulation with the nominal value of the inductor (which is of no significance in a tunable oscillator). 4. There is a discrepancy in higher harmonics, especially at 5th and 6th, we attribute it to the capacitor-coupling effects in the resonator which were not accounted for.

Slide 45 CONCLUSIONS

CONCLUSIONS

- Oscillator Review
- Oscillator Analysis
  - Small Signal
  - Large Signal
  - Oscport

We have presented theoretical background and applications for three methods of oscillator analysis: 1. The local linearization (small signal analysis as we called it) provides oscillations conditions for varying bias. 2. The large signal analysis provides also frequency, power, and harmonic content of the steady state oscillations. Consequently we can predict global behavior of an oscillator including, for example, resonator saturation or the phase noise. 3. The oscport method is easy to apply and also provides frequency, power, and harmonic content of steady state oscillations; moreover it allows to present oscillations as functions of multiple circuit parameters. Some global results, however, like phase noise analysis can be obtained only via harmonic balance analysis. We present three real-life design examples for which our methods are used. For two YIG oscillators, at early design stages, we predict the qualitative global behavior. For the RF oscillator we compare measured and simulated results of the finished circuit.

REFERENCES:

[1] J. E. Brittain, "The Evolution of Electrical and Electronic Engineering..." IEEE Proc., vol.77, no.6, June 1989, pp 837-856.

[2] J.B. Tsui - Microwave Receivers and Related Components, Peninsula Publ. 1985.

[3] A. P. S. Khanna - "Oscillators" Ch.9.in "Microwave Solid State Circuit Design" by. I. Bahl P. Bhartia, J.Wiley 1988.

[4] P. F. Combes, J. Graffeuil, J-F. Sautereau - Microwave Components, Devices, and Active Circuits, J.Wiley 1987.

[5] K. S.Kundert, A. Sangiovanni-Vincentelli - "Simulation of Nonlinear Circuits in the Frequency Domain", IEEE Trans., vol.CAD-5, no.4, October 1986, pp 521-535.

HEWLETT PACKARD

Printed in U.S.A. 4/90
5952-2388