Varactor Measurements and Equivalent Circuits

There have been two reports [1], [2] of the use of series resonance for evaluation of varactors, and another on the impedance locus of coaxially mounted varactors [3]. There can be considerable error in the assumption of a simple resonance, and, since the author has heard of alarming variations in varactor resistance, which may be a consequence of measurement errors, it seems important to draw attention to the pitfalls.

Although it is possible to use an equivalent lumped circuit for a varactor it does not seem adequate to use a simple version except over a limited band of frequencies. Consider a varactor mounted in a coaxial line, Fig. 1. Starting from the plane of the short-circuit (S) there is a nonuniform line up to the plane of the junction, and, if this line is very short, it will behave approximately as an inductance. At the junction (J) there will be pronounced discontinuities, and, including the junction, the equivalent circuit of Fig. 2(a) might be expected to hold over a wide frequency range. (However, the fringing field effects may not be adequately represented by this circuit.) The transforming effect of the nonuniform section from J to plane R can be approximated by capacitance and inductance, so that one version of an equivalent circuit is that shown in Fig. 2(b) in which all elements except the junction are assumed to be loss-free. At low frequencies, the stray capacitance will be $(C_0+C_1+C_f)$ and this is *not* the same as the so-called package capacitance; but, to the extent that both would be affected by changes in the dielectric tube, there will be some correlation between the low-frequency stray capacitance and the package capacitance.

Assuming that the fringing capacitance C_f, in parallel with the junction, is negligible, some features of this equivalent circuit are as follows.

1) At the first series resonance (ω_s), the effective inductance $1/\omega_s^2C_j$ is

$$L_{\text{eff}} \approx \frac{L_2}{1-\omega_s^2L_2C_2}+\frac{L_1}{1-\omega_s^2L_1C_1}.$$

This would not be significantly affected by reversal of the varactor.[1]

2) The resistance at ω_s is $R_{\text{eff}} \approx R(1-\omega_s^2L_1C_1)^2$ and will be affected by reversal unless $L_1C_1=L_2C_2$.
3) The high impedance resonance will be complex, in detail. It will be affected by reversal of the varactor unless $L_1=L_2$ and $C_1=C_2$. The effects of resistive losses associated with the inductances will also be significant.
4) There will be a second series resonance at a very high frequency.

The much simpler circuit used by Roberts [3] does not seem satisfactory, and Fig. 3 shows the results of fitting his data to a circuit similar to Fig. 2, but with C_0 and C_2 omitted. Unfortunately, there appears to be a serious error in phase for at least one of Roberts' low-frequency results, where the ratio of reactances at 2 Gc/s and 3 Gc/s is incompatible with a locus approaching a resonance. Consequently, there is so much uncertainty about the phase accuracy, especially at the higher frequencies, that it seems pointless to attempt fitting the data to a more complex circuit.

The important point is, however, that there may be an error of -20 per cent in the value of varactor resistance (neglecting fringing capacitance), and if the measurement were performed with a lower junction capacitance resonating at 10 Gc/s the error might be -40 per cent.

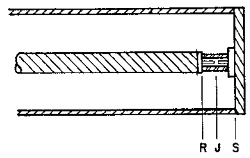

Fig. 1. Coaxial mounting of varactor.

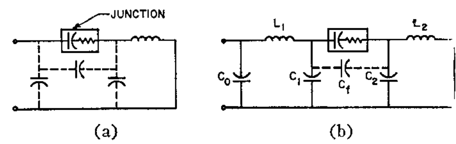

Fig. 2. Equivalent circuits.

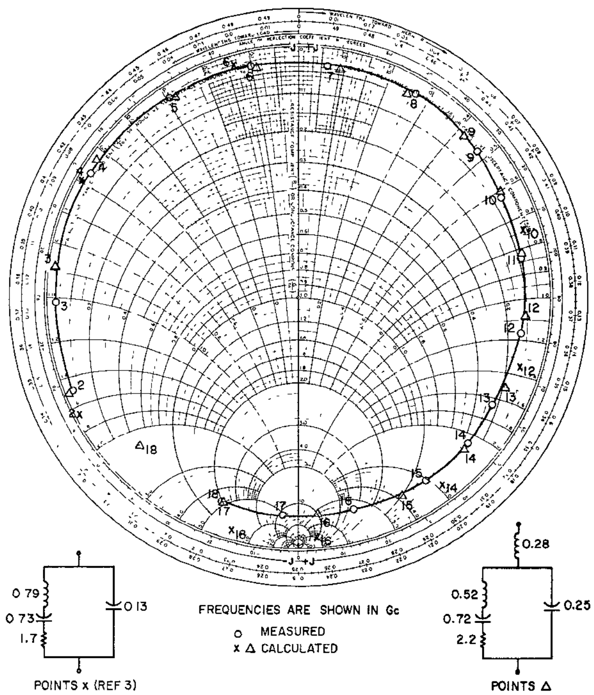

Fig. 3. Comparisons with experimental data.

The correction factor for the resistance may be assessed by varying the first resonant frequency with a constant junction capacitance. Moving the position of the short behind the varactor is one technique for deducing the term $(1-\omega_s^2L_1C_1)^2$.

There are many ways to extend the analysis of an equivalent circuit, and measurement at constant frequency with variable bias should not be neglected. The impedance locus should then fit a constant resistance curve for normal varactors, and the results can be used to check values deduced by other methods. The bias giving minimum reflection coefficient in the region of the first resonance defines the condition corresponding to a resistive mismatch between the source and the varactor resistance, and the locus values of $\Delta X/R$ are not affected by fringing capacitance. Alternatively, by transforming through $\lambda/4$ (approx) an admittance locus lying on a constant conductance curve can be used [2].

With a rectangular waveguide mount, it might be expected that building the diode across the guide would give a good approximation to a simple resonant circuit, as confirmed by DeLoach [1]. He also obtained excellent results in the region of 10 Gc/s with a unit in a Sharpless package. It is suggested that these latter measurements were successful because of the choice of package and mount, and that similar measurements with the more common varactor packages, or with a different arrangement for a Sharpless package, would be unsatisfactory.

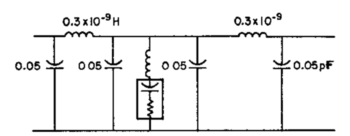

Fig. 4. Suggested equivalent for DeLoach mount.

Each choke joint used with the Sharpless package would give a positive reactance of about 10 ohms at 10 Gc/s. This is estimated from 1.2-mm electrical length in a 40-ohm line. It is suggested that an equivalent circuit of the mount used by DeLoach is approximately that shown in Fig. 4, and that the π sections have an iterative impedance of about 50 ohms, thus

Manuscript received May 7, 1965. The research reported here was sponsored by the Air Force Electronics Systems Div. under Contract AF19(628)2390. Further reproduction is authorized to satisfy the needs of the U. S. Government, and copies will be available as ESD-TR-65-126.

[1] However, it seems probable that C_1 and C_2 will depend upon the position of the short.

matching the guide. (Values shown give $\sqrt{L/C}=55$.) This equivalent circuit cannot be altogether satisfactory and Fig. 10 in [1] clearly shows a change in effective varactor inductance which may be indicative of some transformation of junction impedance.

With the more common types of varactor package, the tolerances are frequently so serious that there would be considerable variations in correction factors necessary to deduce the junction values from the impedance close to resonance.

The use of the series resonant condition does not necessarily make it easier to measure varactors by insertion loss than by reflection methods, although commonly used impedance values are not very suitable for the latter. Blake and Dominick [2] have used the transmission-loss method because of the physical separation of their equipment from the mount, since, at a fixed frequency, measurement of loss is simpler. When the frequency is varied, the changes in generator output, in line losses, and in detector sensitivity complicate the measurement of transmission loss unless a balancing path which includes the reference attenuator is used. In such circumstances, for a limited range of frequencies, there is little or no economy in equipment compared with a reflection technique using two balanced paths connected to a precision directional coupler. Phase information is helpful in detailed analysis, and data such as obtained by Roberts is desirable.

There is need for much further work on evaluation of varactors and especially for the development of methods suitable for routine testing, and which can provide useful data on tolerances in packages and junction characteristics.

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References

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[2] C. Blake and F. Dominick, "Transmission test method for high-Q varactors," *Microwaves*, vol. MTT-4, pp. 18–23, January 1965.
[3] D. A. E. Roberts, "Measurements of varactor diode impedance," *IEEE Trans. on Microwave Theory and Techniques (Correspondence)*, vol. MTT-12, pp. 471–475, July 1964.

Submillimeter Wave Harmonic Mixing

The difference frequency between harmonics of millimeter wave oscillators has been observed at submillimeter wavelengths using a crossed-waveguide harmonic generator[1] as a harmonic mixer.

Figure 1 shows the experimental setup

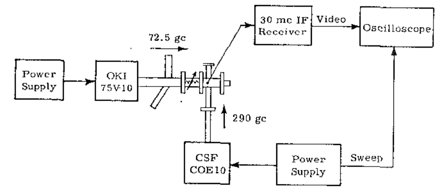

Fig. 1. Harmonics of 72.5-Gc/s klystron mixing with 290-Gc/s carcinotron output and its harmonics.

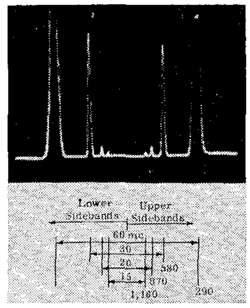

Fig. 2. 1160-Gc/s harmonic mixing.

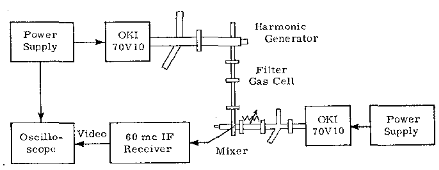

Fig. 3. Harmonic mixing using two 70V10 klystrons.

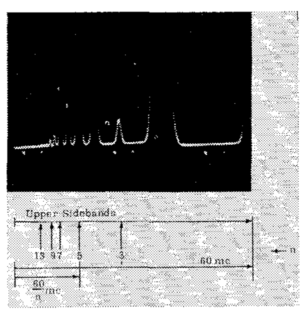

Fig. 4. 950-Gc/s harmonic mixing.

used in mixing the 290-Gc/s output of a carcinotron with the 72.5-Gc/s output of a klystron. The carcinotron was swept about 100 Mc/s and the klystron was operated CW. The difference frequency signals were amplified in a 30-Mc/s IF amplifier and the detected video output was displayed on an oscilloscope. Figure 2 shows the receiver output. Upper and lower sidebands are displayed. The nth harmonic of the carcinotron mixes with the 4 nth harmonic of the klystron to produce IF signals. The upper and lower sidebands are separated by $60/n$ Mc/s on the carcinotron sweep. The highest harmonic observed was the fourth at 1160 Gc/s.

Figure 3 shows the experimental arrangement used in mixing the output from a 72.9-Gc/s klystron with harmonics of a second 72.9-Gc/s klystron. In this case, one crossed-waveguide device was used as a harmonic generator and an identical unit was used as a harmonic mixer. Figure 4 shows the upper sidebands of the receiver output. Thirteen harmonics were observed. A narrow-band receiver is needed to resolve adjacent harmonics. For example, the tenth and twelfth harmonics are separated by 1 Mc/s at the fundamental with a 60-Mc/s IF. In order to determine how many of these harmonics were generated in the multiplier and propagated to the mixer and how many were generated in the mixer, a waveguide filter cutting off the second harmonic was inserted between the multiplier and the mixer. Only the third through sixth harmonics were observed. These were shown to be generated by the multiplier by observing absorptions in a gas cell with carbonyl sulfide, OCS. The signals corresponding to $n=1$ and $n=7$ through 13 were generated in the mixer.

In another experiment, both 72.9-Gc/s klystron outputs were fed in the RG-98 waveguide input of a mixer. One tube was connected to the regular input and the tuning short was removed to accept the second input. In this case so many harmonic beats were observed that the higher harmonics were not resolved. More than 20 harmonics of the 72.9 input were observed.

Harmonic mixing experiments similar to these have been reported by Murai[2] who observed beats as high as 750 Gc/s using a IN53 crystal. A millimeter wave superhetrodyne system using similar techniques was used by Johnson[3] for spectroscopy in the 100- to 150-Gc/s region.

Since 70-Gc/s klystrons can be phase locked to crystal oscillator harmonics, this harmonic mixing technique can be used for accurate measurement of the frequency of submillimeter oscillators (far-infrared lasers) and for phase (or frequency) stabilization of these sources.

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Manuscript received June 21, 1965.

[1] Devices of this type were first used by spectroscopists to generate millimeter waves from centimeter klystrons. For example, see W. C. King and W. Gordy, *Phys. Rev.*, vol. 93, p. 407, 1954. The units used in these experiments have RG-98 and RG-135 waveguides.

[2] A. Murai, "Heterodyne beat between submillimeter components generated in a crystal detector," presented at the 1964 Internat'l Conf. on Microwaves, Cur ent Theory, and Information Theory, Tokyo, Japan.

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Magnetostriction Effects in Remanence Phase Shifters

One type of remanence phase shifter[1] consists of a microwave ferrite toroid located in a waveguide. Close mechanical fit between ferrite and waveguide is desirable to eliminate reflection spikes, and to provide an adequate thermal path. Such structures typically develop mechanical pressure on the ferrite, and this pressure may vary with temperature, due to the unequal expansion of the waveguide and ferrite with temperature.

Manuscript received May 25, 1965.

[1] L. Levey and L. Silber, "A fast switching X-band circulator utilizing ferrite toroids," *1950 IRE Wescon Conv. Rec.*, pt. 1, pp. 11–20.