Some rules for the choice of the C(V) characteristic for the design of frequency triplers with symmetrical varactors

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***Abstract* - In this paper we consider multipliers designed with varactors that have a symmetric *C(V)* capacitance-voltage characteristic, i.e. triplers, quintuplers, ... We show that for a tripler the optimal *C(V)* characteristic is not the most abrupt one, as stated in much works, but rather a cosine-like one. Our work is validated with the design of a frequency tripler based on the use of HBVs non-linear transmission lines. We obtained a significant improvement for the maximum conversion efficiency when a cosine *C(V)* is used instead of an abrupt one, for a 15 HBVs NLTL frequency tripler.**

I. INTRODUCTION

In this paper we consider frequency triplers realized with varactors that have a symmetric *C(V)* capacitance-voltage characteristic and our results are applied to the design of an heterostructure barrier varactor's (HBV) based non-linear transmission line (NLTL).

Some encouraging results have been published in the past concerning varactor frequency multipliers based on NLTLs. Rodwell demonstrated a monolithic V band doubler with 22% maximum conversion efficiency, and a W band tripler with 7.6% conversion efficiency [1]. More recently, results were published of an hybrid tripler with 10dBm output power at 130.5GHz, 7% maximum conversion efficiency. The multiplier consisted in a finline section periodically loaded by 15 HBVs [2]. Also encouraging results have been achieved by IEMN with a monolithic coplanar transmission lines loaded by 8 HBV as a tripler @ 60GHz [3]. In the design of varactor frequency multipliers, some design rules seem to be assumed by a lot of authors. Particularly, a maximum slope of *C(V)* around zero-volt bias is usually pointed out [4]. However, in the case of high harmonics, it is usually better to cascade low-harmonic multipliers, because of losses and the difficulties in realizing a large number of idlers [5]. Inversely, in the case of low-order multipliers (doublers and triplers), the question has to be addressed about the necessity to use varactors with *C(V)* having a zero-volt bias maximum slope. They generate many high-order harmonics, which poorly contribute to low-order harmonic generation when losses occur. Moreover, today, new components like HBVs offer the possibility to easily change the shape of *C(V)* while with Schottky varactors we could only change the doping profile.

So, ***is the zero-volt bias maximum slope an optimum for C(V) or could we find a better C(V) shape for the design of a tripler with HBVs ?***

Next, if we have to optimize this shape, ***which varactor capacitance model will be taken into account for the design ?*** Today, the varactor cut-off frequency, define by (1), is the most used one [3]. Although for Schottky diode or for the new varactors as HBV diodes.

$$f_{cd} = (S_{max} - S_{min})/2\pi R_s \qquad (1)$$

with $S_{max} = 1/C_{min}$ and $S_{min} = 1/C_{max}$ the minimum and maximum elastances, and R_s the varactor series resistance.

In this paper, we first show why for HBVs, relation (1) does not constitute the most appropriate choice to calculate the varactor cut-off frequency. Next we show that a cosine *C(V)* shape gives better results than a typical zero-volts bias maximum slope *C(V)* characteristic. And finally the new concepts developed are applied to the design of a 15 HBVs frequency tripler based on NLTLs for validation.

II. VARACTOR CUT-OFF FREQUENCY

In the literature, we can find two different approaches for the design of multipliers. On one hand, the varactor choice is based on the cut-off frequency f_{cd} given by (1). In that case, an equivalent cut-off capacitance is defined by (2):

$$C_{eq} = 1/(S_{max} - S_{min}) \qquad (2)$$

On the other hand designers use the concept of the varactor large signal equivalent capacitance (3):

$$C_{ls} = \frac{1}{V_{max} - V_{min}} \int_{V_{min}}^{V_{max}} C(v).dv \qquad (3)$$

with the associated large-signal cut-off frequency (4):

$$f_{c,ls} = 1/(2\pi R_s C_{ls}) \qquad (4)$$

A. C_{eq} and C_{ls} versus the voltage range

When Schottky barrier varactors are considered, the *C(V)* characteristic is (5):

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$$C(v) = C_{j0} / (1 - v/\phi)^{\gamma} \quad (5)$$

So the two normalised capacitances expressions versus the amplitude voltage are (6) and (7):

$$\frac{C_{eq}}{C_{j0}} = 1 / \left[(1 + V_m / \phi)^{\gamma} - 1 \right] \quad . \quad (6)$$

$$\frac{C_{ls}}{C_{j0}} = \frac{\phi}{V_m.(1-\gamma)} \left(\left(1 + \frac{V_m}{\phi} \right)^{-\gamma+1} - 1 \right) \quad (7)$$

In the case of HBV, we consider the $C(V)$ of those fabricated and measured by IEMN [6].

Fig. 1 compares the $C(V)$ characteristics of a GaAs Schottky diode with $\phi = 0.75V$ and $\gamma = 0.5$ (graded doping profile) and those of IEMN [6]. The normalised capacitances are shown in Fig. 2.

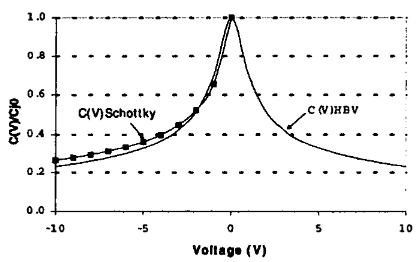

Fig. 1. HBV and Schottky diode $C(V)$ characteristics.

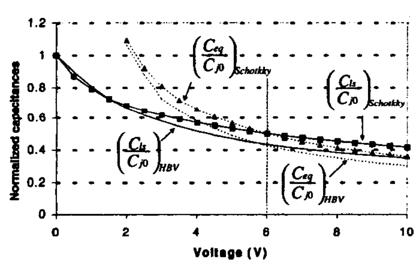

Fig. 2. Normalised capacitances versus voltage amplitude.

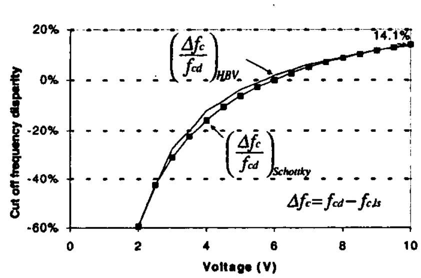

Fig. 3. Relative difference on f_{cd} for HBV and Schottky diode.

The cut-off frequency f_{cd} is usually defined for one standard voltage range V_{cmax}=6V [7]. We can see in Fig. 2 that the normalised C_{ls} and C_{eq} are equal only for this voltage. Fig. 3 shows the relative difference between the two cut-off frequencies with R_s=5Ω. For 10V, it reaches 14.1%. It is evident that these two cut-off frequency models are different, but ***what is the best cut-off frequency definition for the multipliers design ?*** We will compare these two approaches for an actual multiplier design case and try to answer this question.

B. Dynamic model choice

First, note that whatever model is used (C_{eq} or C_{ls}), the cut-off frequency is calculated from an approximation of the varactor behavior, because we use small-signal parameters to predict the behavior of a non linear device fed by a large signal generator. We will compare these small-signal models to the large-signal model taken as a reference by running simulations to find the best choice between C_{eq} and C_{ls}. The simple simulation circuit is described in Fig. 4. SPICE is used for all the simulations. The generator used to feed the varactor is a square voltage varying between $\pm V_{cmax} = \pm 10V$ so that a high-harmonics content signal is applied to the varactor.

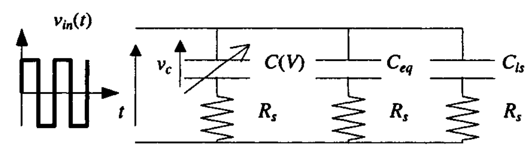

Fig. 4. Electrical models comparison for HBV.

In Fig. 5, simulation results show that the relative error between the harmonics contained in v_c for C_{eq} and C_{ls}, and the harmonics obtained for the large-signal model.

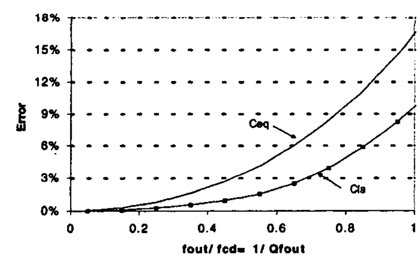

Fig. 5. Relative error between small/large-signal approaches.

This error increase versus $1/Q_{fout}$, where

$$Q_{f_{out}} = f_{cd} / f_{out} \quad (8)$$

f_{out} is the output multiplier frequency.

In any case the relative error due to C_{eq} is higher than the C_{ls} one. Thus C_{ls} approach will be preferred.

We can see in Fig. 3 that f_{cd} is overvalued when V>6V. When (1) is used, the varactor voltage-variable capacitance choice is more constraining than using (4).

III. Varactor C(V) Shape

A. Harmonics Generation

First we can remember how the harmonics are generated by a varactor. Let consider a varactor embedded between two transmission lines of characteristic impedance Z_0, loaded by Z_c. Fig. 6 shows the simplified electrical model used for this purpose (with: $Z_0=\sqrt{L_0/C_0}$). The model is valid for frequencies for below the Bragg frequency: $f_B=1/\left(\pi\sqrt{L_0C_0}\right)$.

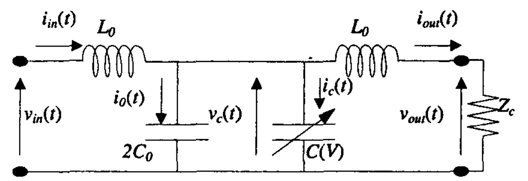

Fig. 6. Simplified equivalent electrical circuit used for the study of the harmonics generation in HBVs.

$$\begin{aligned} v_{out}(t)-v_{in}(t) &= -L_0\frac{d}{dt}\left(i_0+2i_{out}+i_c\right) \\ &= -L_0\frac{d}{dt}\left(2C_0\frac{dv_c}{dt}+\frac{2}{Z_c}v_{out}+C(v_c)\frac{dv_c}{dt}\right) \end{aligned} \tag{9}$$

The first two terms are linear. The third term is non-linear and depends on $v_c(t)$. Therefore, it is clear that the output harmonics on $v_{out}(t)$ are generated by the varactor current harmonics. That's why the optimal C(V) shape can be chosen by studying the mechanisms involved in the varactor current harmonic generation. Now, we will show the limitations involved by a classical HBV shape with zero-volt bias maximum *C(V)* slope.

B. Classical HBV shape harmonics generation

When the HBV (*C(V)* characteristic of Fig. 1) is fed by a 30dBm power sinewave (or 10V on 50Ω)) fast transitions occur in the waveforms of $i_c(t)$ (Fig. 8) and *C(t)* (Fig. 10) around zero-volt (time about 12 ps and 38 ps). At this voltage, *C(V)* slope and the sinewave variations are simultaneously maximum. These fast transitions generate high-order harmonics in $i_c(t)$ and *C(t)* associated spectrum (Fig. 9). This generation spreads the energy over a wide spectrum. It should not be the best way to design a tripler. In theory, when losses are neglected, high harmonics would recombine together to reach 100% conversion efficiency [5]. Actually, the conversion efficiency will not be maximum with this *C(V)* shape because the high-order harmonics energy will be partly lost.

C. Determination of the optimal C(V)

To minimize the generation of high-order harmonics and reinforce the third harmonic generation, the idea is to use a *C(V)* characteristic with a small slope around zero volt. A simple cosine shape seems to correspond to this requirement. To point out the interest of taking a cosine *C(V)*, we compare its shape to three other one: the HBV's *C(V)* shape described in Fig. 1 (obtained from a third order expression [8]), a first and second order polynomial *C(V)* characteristic. The four shapes are plotted in Fig. 7 for the case $k=C_{min}/C_{max}=0.231$ and $V_{cmax}=10V$.

These *C(V)* expressions are defined as follows:

$$C(v_c)=C_{j0}\left(1-(k-1)|v_c|/v_{cmax}\right) \tag{10}$$

$$C(v_c)=C_{j0}\left(1+\alpha|v_c|/v_{cmax}+\beta(v_c/v_{cmax})^2\right) \tag{11}$$

with : $\alpha=-0.165$ and $\beta=0.079$.

$$C(v_c)=C_{j0}\left(0.5(k+1)\left\{1+cos\left(\pi|v_c|/v_{cmax}\right)\right\}\right) \tag{12}$$

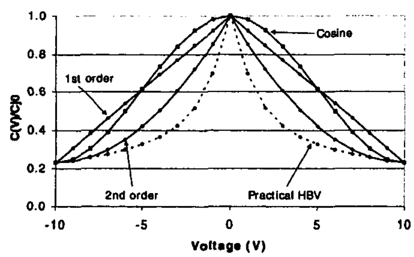

Fig. 7. The four shapes $C(v_c)/C_{j0}$ versus v_c.

All the *C(V)* characteristics except the cosine one have at zero-volt bias a dC/dv_c discontinuity.

Fig. 8 shows the waveforms obtained for $i_c(t)$ (current through the varactors) and Fig. 9, its associated spectrum, when the varactors are fed by a sinewave :

$$v_{in}(t)=V_{in}\sin(\omega_{in}t) \tag{13}$$

with $f_{in}=\omega_{in}/2\pi=20GHz$ and $V_{in}=10V$.

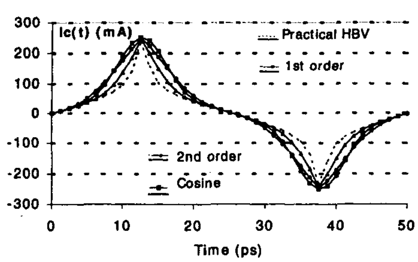

Fig. 8. Waveform of the current flowing through the varactors

These results are interesting. We can see that with the cosine shape, the maximum conversion efficiency is obtained for the third harmonic, and the harmonics magnitude decreases rapidly and reaches zero after the seventh harmonic. As higher the order polynomial model is, as wide the spectrum is, but for a multiplier only one frequency is required.

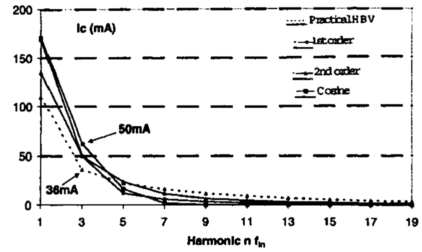

Fig. 9. Waveform of the current flowing through the varactors and the associated spectrum.

Fig. 10 shows the four $C(v_c(t))$ waveforms. The cosine shape gives the $C(v_c(t))$ closest to the optimum capacitance form:

$$C(t) = C_{ls} + C_{opt}\sin(\omega_{in}t + \varphi_{opt}) \tag{14}$$

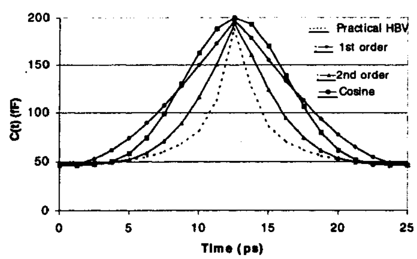

Fig. 10. $C(v_c(t))$ waveforms obtained for the 4 shapes.

We define the optimal capacitance form for which only two frequencies will be generated, the fundamental and the third harmonic:

$$i_c(t)=C(v_c(t))\frac{dv_c}{dt} \tag{15}$$

$$i_c(t)=C_{ls}.\cos(\omega_{in}.t)+\frac{C_{opt}}{2}[\sin(\omega_{in}.t+\varphi_{opt})+\sin(3\omega_{in}.t+\varphi_{opt})]$$

IV. APPLICATION

We have applied these concepts for the design of a NLTL tripler with a cosine $C(V)$ shape. The NLTL was designed with a coplanar waveguide (CPW) transmission line on an InP substrate. HBV and CPW metallic losses were not taken into account in the simulations. The NLTL has been designed as follows :

1. Choice of the CPW characteristic impedance Z_{cpw} and the Bragg frequency f_B (between the third and fifth harmonics),
2. Calculation of the large signal varactor capacitance C_{ls},
3. Optimization of f_B and Z_{cpw} by running SPICE simulations to obtain the maximum conversion efficiency.

Fig. 11 shows the results obtained for the shape of Fig. 1 and the cosine one. The use of the cosine shape for the HBV clearly improves the conversion efficiency. The improvement is about 10% for 15 HBVs.

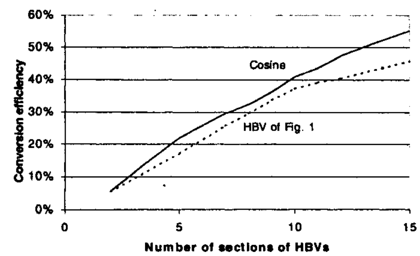

Fig. 11. Conversion efficiency on the third harmonic for the NLTL tripler, versus the number of sections N.

V. CONCLUSION

A study of symmetrical $C(V)$ varactors optimal shape for a maximum conversion efficiency has been carried out. For tripler frequency multipliers we show that a shape with a maximum slope around zero-volt bias was not optimal. An application on a NLTL tripler has been designed. With a cosine $C(V)$ shape, the converted energy can be concentrated in the 3rd harmonic. The cosine shape use improves the conversion efficiency of about 10%.

REFERENCES

[1] E. Carman & al., " V-band and W-band broadband, monolithic distributed frequency multipliers, " *1992 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, pp. 819-822, June 1992.
[2] S. Hollung & al., " A distributed Heterostructure Barrier Varactor Frequency Tripler, " *IEEE Mic. & Guided Wave Lett.*, vol. 10, no. 1, pp. 24-26, Jan. 2000.
[3] M. Fernandez, E. Delos, X. Mélique, S. Arscott and D. Lippens , "Monotithic coplanar transmission tines loaded by heterostructures barrier varactors for a 60GHz," *Microwave and Wirelress Components Letters*, pp. 498 December 2001.
[4] L. Dillner & J. Stake and E. L. Kollberg, "Analysis of symmetric-varactor frequency multipliers" *Microwave and Optical Technology Letters* , vol. 15, n°1, pp 26-29 May 1997.
[5] S.A. Maas, *The RF and microwave circuit design cookbook*, Boston: Artech House, 1998.
[6] X. Mélique & al.,"Fabrication and performance of InP-Based HBV in a 250GHz Waveguide Tripler", *IEEE Trans. on Mic. Theory & Tech.*, vol. 48, no. 6, pp. 1000-1006, June 2000.
[7] S.A. Maas, *Non linear microwave circuits*, pp. 290-319, NY Boston: IEEE press, 1996.
[8] K. Krishnamurthi & R.G. Harrison, "Analysis of symmetric-varactor frequency triplers, " *1993 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, pp. 649-652, June 1993.