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The Overdriven Varactor Upper Sideband Upconverter

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Abstract—**The equations for the overdriven upper sideband upconverter are derived and computer solutions are given for the abrupt junction, graded junction, and punch through varactor. The necessary design parameters are presented for the design of an upconverter. The performance of the abrupt, graded, and punch through varactors are compared.**

LIST OF SYMBOLS

$C(V)$ = Junction capacitance as a function of voltage across it in the reverse direction

Eff = Efficiency—power out divided by total power in

$f_c(V)$ = Cutoff frequency [defined by (16)] as a function of V

P_K = Total power into varactor for $K=1$, and 2 and out of varactor for $K=3$ [see (12)]

P_{0K} = The power into the lossless varactor at frequency ω_K ($K=1, 2, 3$)

P_{DK} = The power dissipated in R_s at frequency ω_K ($K=1, 2, 3$)

q = The charge on the varactor

$\bar{q}$ = The normalized charge defined by (2)

q_ϕ = The charge due to the contact potential

Q_0 = Average normalized charge on the varactor

$\overline{Q}_K$ = Normalized charge on varactor at frequency ω_K ($K=1, 2, 3$) defined by (4)

Q_B = Charge at breakdown

R_K = Real part of "impedances" (ratio of voltage to current) across the varactor at ω_K

$\overline{R}_K$ = Normalized resistance defined by (19)

R_S = Parasitic series resistance of diode

$S_{\max}$ = Maximum value of the varactor elastance—value of elastance at breakdown

T = Period of charge waveform

V_B = Breakdown voltage

V = Voltage across lossless varactor in reverse direction

$\overline{V}$ = Normalized voltage across varactor defined by (2)

$\overline{V}_{KC}$ = Voltage across diode at frequency ω_K in phase with the current at ω_K

$\overline{V}_{KS}$ = Voltage across diode at frequency ω_K out of phase with current at ω_K

V_0 = dc voltage across varactor

$\overline{V}_0$ = Normalized dc voltage across varactor

X = Parameter in (15)

X_K = Imaginary part of "impedances" (ratio of voltage to current) across the varactor at ω_K

$\overline{X}_K$ = Normalized imaginary part of the impedance defined by (19)

α = Loss coefficient defined by (17)

β = Normalized output power defined by (18)

γ = Varactor law defined by (1)

ϕ = Contact potential

$\omega_1 = 2\pi$ times the input frequency

$\omega_2 = 2\pi$ times the pump frequency

$\omega_3 = 2\pi$ times the output frequency

$\omega_c = 2\pi$ times $f_c(V_B)$

θ_K = The phase angle of the charge waveform at ω_K for $K=2$ and 3 in (4)

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Introduction

THE UPPER sideband upconverter (USBUC) is used in many communication systems where signals are amplified at lower frequencies and then translated to higher frequencies. Many of these applications require high efficiency and large power outputs. There is an inverse relationship between cutoff frequency of a varactor and power handling capacity.[1] Since overdriving[2] the varactor USBUC increases the power handling capability with little change in the efficiency; it is a practice which is commonly employed.

In this paper the equations for the overdriven USBUC are derived and computer solutions are given for the abrupt junction, graded junction, and punch through varactor. The necessary parameters are given for the design of up-converters at desirable operating points.

Curves of the efficiency and power output for zero average charge as a function of the ratio of the output to input frequency are given for the abrupt junction, graded junction, and punch through varactor. Curves are also given comparing the performance of the three types of varactors.

Previous work in the varactor USBUC include a number of analyses of the non-overdriven abrupt junction varactor USBUC[3]–[5] and a limited amount of work in the graded junction non-overdriven USBUC.[6] The only works in the literature on the overdriven USBUC is a proof[2] that the abrupt junction USBUC when overdriven such that the average charge equals zero handles four times the power at the same efficiency and impedance level as the fully driven[1] USBUC and design parameters[7] for the punch through varactor [$\gamma=0$ in (1)].[2]

In this paper we shall assume that the voltage never exceeds breakdown.[3] The normalized average charge Q_0 is allowed to assume values between 0.5 and 0. The value 0.5 corresponds to the fully driven[1] (but not overdriven) case, while for $Q_0<0.5$ the varactor is in forward conduction for part of the cycle. $Q_0=0$ implies that the diode is in forward conduction for half the cycle. $Q_0<0$ implies that there is average stored charge in the forward direction. Recombination must then occur and a dc current equal to Q_0/τ, where τ is the recombination time, will then flow. This case was investigated by the computer program and it was found that the power output could be increased but that the efficiency decreased rapidly. The range of interest was, therefore, confined to $0\leq Q_0\leq 0.5$. It was assumed over this region that the recombination current was negligible.

[1] The varactor is fully driven when the voltage across the diode varies from V_B to $-\phi$. The average normalized charge Q_0 (4) is then 0.5.

[2] The results in Grayzel[7] were obtained from the computer program described in this paper.

[3] It is possible to operate a multiplier or USBUC such that the varactor goes into avalanche for part of the cycle. Rectification will then occur and hence power will be dissipated at dc. There will also be noise generated by the avalanche process and a tendency for instability. The power handling capability, however, can be increased using this technique. There are no analyses in the literature for this case.

Procedure

The varactor will be approximated by a lossless nonlinear capacitor in series with a fixed resistance R_S. It is well known[8] that the voltage charge relationship for an ideal lossless varactor in the reverse direction is given by

$$\bar{V} = \bar{q}^{1/1-\gamma} \tag{1}$$

where

$$\bar{V} = \frac{V+\phi}{V_B+\phi} \tag{2a}$$

$$\bar{q} = \frac{q+q_\phi}{Q_B+q_\phi} \tag{2b}$$

and where $\bar{V}$ and $\bar{q}$ are by definition positive when the varactor is reverse biased. When $\bar{q}$ is negative the varactor is in forward conduction. The elastance is then zero and the junction can be approximated by a short circuit. The voltage charge relationship is then[4]

$$\begin{aligned} \bar{V} &= \bar{q}^{1/1-\gamma} \qquad \bar{q}>0 \\ &= 0 \qquad\qquad \bar{q}<0 \end{aligned} \tag{3}$$

$\gamma=\frac{1}{2}$ corresponds to the abrupt junction, $\gamma=\frac{1}{3}$ the graded junction, and $\gamma=0$ the punch through varactor.

It is assumed that current flows through the varactor at only three frequencies ω_1 the signal frequency, ω_2 the pump frequency, and $\omega_3=\omega_1+\omega_2$ the output frequency. The normalized charge $\bar{q}$ defined by (2b) can be written in general

$$\bar{q} = Q_0 + 2\bar{Q}_1 \sin(\omega_1 t) + 2\bar{Q}_2 \sin(\omega_2 t+\theta_2) + 2\bar{Q}_3 \sin(\omega_3 t+\theta_3). \tag{4}$$

For a given charge waveform the voltage can be determined from (1). The Fourier coefficients of the voltage can then be found which yields the voltages at ω_1, ω_2, and ω_3. The current through the varactor is determined by differentiating (4). Once the voltage and current at each frequency is known the powers and impedances across the lossless variable capacitance can be computed. Computation of the loss in R_S at each frequency which equals $2(Q_B+q_\phi)^2\ \bar{Q}_K^2\ \omega_K^2 R_S$ then enables the determination of the efficiency.

The normalized voltage across the lossless varactor at ω_K in phase with its respective current is given by

$$\bar{V}_{KC} = \lim_{T\to\infty} \frac{2}{T}\int_{-T/2}^{T/2} \bar{V}\cos(\omega_K t+\theta_K)dt. \tag{5}$$

If ω_1, ω_2, and ω_3 are commensurate with period T then

$$\bar{V}_{KC} = \frac{2}{T}\int_0^T \bar{V}\cos(\omega_K t+\theta_K)dt. \tag{6}$$

[4] This model has been used by many workers in analyzing various varactor multiplier circuits.[9],[10] Excellent agreement has been found experimentally for these cases.

The voltage at ω_K out of phase with its respective current is given by

$$\overline{V}_{KS} = \frac{2}{T}\int_0^T \overline{V} \sin(\omega_K t + \theta_K) dt \tag{7}$$

and the normalized dc voltage across the diode is given by

$$\overline{V}_0 = \frac{1}{T}\int_0^T \overline{V} dt. \tag{8}$$

These integrals can be evaluated by numerical methods. At n equally spaced points in the interval from 0 to T $\bar{q}$ is evaluated (4), $\overline{V}$ is then determined from (3) and the integrand is then evaluated. A weighted sum (given by Simpson's rule) of the values of the integrand at the n points gives the value of the integral. Commensurate frequencies are chosen so that T is finite; however, the sets of frequencies chosen must not be harmonically related. For instance, if $\omega_2 = 3\omega_1$ then power will be converted to $\omega_3 = 4\omega_1$ not only by the upconversion process but also by the quadrupling of the input signal, and hence one gets an erroneous answer. If one, on the other hand, chooses $2\omega_2 = 3\omega_1$ then $2\omega_3 = 5\omega_1$ and harmonics of ω_1 and ω_2 do not coincide with ω_3.

If the varactor is resonated at the output frequency then $\theta_2 = \theta_3$.[5] θ_2 represents the phase between the input and pump signal which generally can not be controlled. However, for a given set of charge coefficients Q_0, $\overline{Q}_1$, $\overline{Q}_2$, and $\overline{Q}_3$ the efficiency and power output are independent of θ_2 a fact which was verified on the computer. θ_2 and θ_3 in (4) were, therefore, set equal to zero for convenience.

The output power from the lossless nonlinear capacitor is given by

$$\begin{aligned} P_{0K} &= \omega_K \overline{Q}_K \overline{V}_{KC}(V_B + \phi)(Q_B + q_\phi) \\ &= \frac{(V_B + \phi)^2 \omega_K}{(1 - \gamma) S_{\max}} \overline{Q}_K \overline{V}_{KC} \end{aligned} \tag{9}$$

where the relationship[11]

$$Q_B + q_\phi = \frac{V_B + \phi}{(1 - \gamma) S_{\max}} \tag{10}$$

was used. The power dissipated in R_S at frequency ω_K is given by

$$\begin{aligned} P_{DK} &= 2R_S \omega_K^2 \overline{Q}_K^2 (Q_B + q_\phi)^2 \\ &= \frac{(V_B + \phi)^2}{(1 - \gamma)^2} \frac{\omega_K^2}{S^2_{\max}} \overline{Q}_K^2 R_S. \end{aligned} \tag{11}$$

The total powers into the varactor at ω_1 and at ω_2 and out of the varactor at ω_3 are given by[6]

$$\begin{aligned} P_1 &= P_{01} + P_{D1} \\ P_2 &= P_{02} + P_{D2} \\ P_3 &= -P_{03} - P_{D3}. \end{aligned} \tag{12}$$

The efficiency is given by

$$\text{Eff} = \frac{P_3}{P_1 + P_2} \cdot \tag{13}$$

The real and imaginary values of the "impedances" at ω_K are given by[7]

$$\begin{aligned} R_K &= R_S + \frac{V_B + \phi}{(Q_B + q_\phi)\omega_K} \frac{\overline{V}_{KC}}{2\overline{Q}_K} \\ &= R_S + (1 - \gamma) \frac{S_{\max}}{\omega_K} \frac{\overline{V}_{KC}}{2\overline{Q}_K} \end{aligned} \tag{14a}$$

and

$$X_K = \frac{(1 - \gamma)}{2} \frac{S_{\max}}{\omega_K} \frac{\overline{V}_{KS}}{\overline{Q}_K} \cdot \tag{14b}$$

Thus given Q_0, $\overline{Q}_1$, $\overline{Q}_2$, and $\overline{Q}_3$ the operating point of the varactor is completely determined. The problem that remains is for a given Q_0 to find $\overline{Q}_1$, $\overline{Q}_2$, $\overline{Q}_3$ subject to the constraint $\bar{q} \leq 1$, which maximizes the output power or the efficiency.

Nelson[12] has derived the constraint on Q_0, $\overline{Q}_1$, $\overline{Q}_2$, and $\overline{Q}_3$ such that $\bar{q} \leq 1$ for the non-overdriven case. This constraint can be generalized for the overdriven case by introducing arbitrary Q_0. The constraint then becomes

$$Q_0 + \overline{Q}_1 \cos X + \sqrt{\overline{Q}_2^2 + \overline{Q}_3^2 + 2\overline{Q}_3\overline{Q}_2 \sin X} \leq 1 \tag{15a}$$

where

$$\sin X = \frac{\overline{Q}_2 \overline{Q}_3 \cos X}{\overline{Q}_1 \sqrt{\overline{Q}_2^2 + \overline{Q}_3^2 + 2\overline{Q}_2\overline{Q}_3 \sin X}} \cdot \tag{15b}$$

When inequality (15a) is satisfied with the equal sign we are on the breakdown curve. Increasing any of the charge coefficients will cause the maximum voltage to exceed V_B and breakdown will occur. The maximum power point always lies on the breakdown curve.[8] since if one postulated a maximum power point which was not on the breakdown curve one could multiply the corresponding $\bar{q}$ (4) by a constant such that the new $\bar{q}$ would lie on the breakdown curve. From (3), $\overline{V}$ would also increase by a constant factor and hence, the power at each frequency would be increased violating our assumption. The maximum efficiency point will also lie on the breakdown curve for $\gamma > 0$ while for $\gamma = 0$ the efficiency is independent of the drive level. This can be

[5] Values of θ_3 not equal to θ_2 were used in the program and it was verified that maximum power output and maximum efficiency occur when $\theta_3 \sim \theta_2$.

[6] Since P_{03} represents the power into the lossless varactor $-P_{03}$ is the power out.

[7] R_3 and X_3 are negative as defined. The load impedance is equal to $-(R_3 + jX_3)$

[8] Assuming $V \leq V_B$.

understood as follows. The cutoff frequency of a varactor is given by

$$f_c(V) = \frac{1}{2\pi R_S C(V)} \cdot \tag{16}$$

For $\gamma > 0$, $C(V)$ decreases for increasing V. Hence, one gets the largest cutoff frequency when V is at its maximum allowable value, i.e., $V = V_B$. Since the efficiency is a monotonic increasing function of the cutoff frequency evaluated at the maximum value of V [see (17)] one gets the highest efficiency when the varactor is driven to V_B. When $\gamma = 0$, $C(V)$ and $f_c(V)$ are both independent of V and, hence, the efficiency is not a function of the drive level. Nevertheless, for $\gamma = 0$ one gets the highest output power for a given efficiency on the breakdown curve.

Thus, to find the maximum power point or the maximum efficiency point for a given Q_0 one can search the Q_1 $\overline{Q}_2$ plane, using (15) to determine $\overline{Q}_3$ for each point. This was done on a 360 computer using a "hill climbing" procedure to find the maximum efficiency point and the maximum power point.

Results

The complete results of the computer analysis are by necessity quite voluminous. Complete design parameters must include the efficiency powers at each of the three frequencies, the real and imaginary parts of the impedances at each of the three frequencies, and the bias voltage. These must be given for different ratios of ω_3/ω_1, for different ratios of ω_c/ω_3, and for different values of Q_0. The complete tables are given in Grayzel.[13] Here we shall extract that data which will allow design of upconverters at desirable operating points.

If $\omega_3 \ll \omega_c$ one can approximate the efficiency by

$$\text{Eff} = e^{-\alpha(\omega_3/\omega_c)}. \tag{17}$$

α was evaluated for $Q_0 = 0$ with $\omega_c = 10^4\ \omega_3$ and is plotted at the maximum efficiency point and the maximum power output point *VS* ω_3/ω_1 in Figs. 1 through 3 for $\gamma = \frac{1}{2}, \frac{1}{3}$, and 0, respectively. These values of α, when substituted into (17), give values of efficiency which are lower than the actual efficiencies computed by the program by less than 3 percent for Eff>0.75 and less than 5 percent for Eff>0.35.

Figs. 1 through 3 also give the value of β at the maximum efficiency and the maximum power point (for $Q_0 = 0$) where β is related to the output power P_3 by

$$P_3 = \frac{(V_B + \phi)^2}{S_{\max}}\ \omega_3\beta \quad \text{(milliwatts).} \tag{18}$$

β was calculated for $\omega_c = 10^4\ \omega_3$. It can be seen from these figures that for large values of ω_3/ω_1 at the maximum efficiency point the power handling capability is low. This can be understood as follows. The maximum power output point lies near the point $\overline{Q}_1 = \overline{Q}_2 = \overline{Q}_3$[5] while the maximum efficiency point lies near the point $\omega_1\overline{Q}_1^2 = \omega_2\overline{Q}_2^2 = \omega_3\overline{Q}_3^2$.[6] For large values of ω_3/ω_1 these points lie far apart. There is, therefore, a trade-off between efficiency and power output.

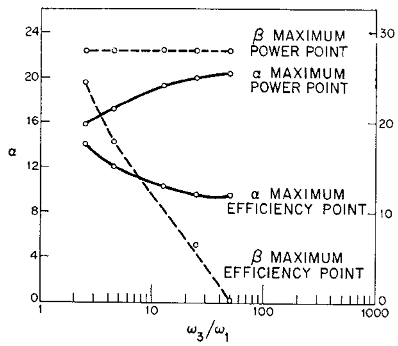

Fig. 1. $\gamma = \frac{1}{2}$, α and β versus ω_3/ω_1 at the maximum efficiency point and at the maximum power point.

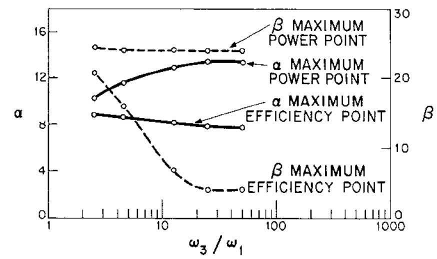

Fig. 2. $\gamma = \frac{1}{3}$, α and β versus ω_3/ω_1 at the maximum efficiency point and at the maximum power point.

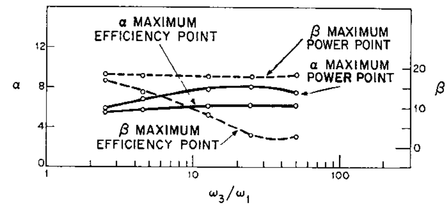

Fig. 3. $\gamma = 0$, α and β versus ω_3/ω_1 at the maximum efficiency point and at the maximum power point.

A typical curve is given in Grayzel.[5] A point close to the maximum power point represents the best choice for most applications. We shall, therefore, only give design parameters for the maximum power point. As mentioned, the value of β in Figs. 1 through 3 are calculated for $\omega_c/\omega_3 = 10^4$. As ω_c/ω_3 decreases both the efficiency and the power output decrease. It has been found empirically that a good estimate of the actual output power can be obtained by multiplying the value of β given in Figs. 1 through 3 by $(1+2\ \text{Eff})/3$. The value of P_3 calculated by (18) is then within 5 percent of the actual value for Eff>0.50.

It was found that for $\gamma = \frac{1}{3}$ and $\gamma = 0$ one gets greatest efficiency and power handling capability for $Q_0 = 0$. For $\gamma = \frac{1}{2}$, however, one can get slightly better efficiency for $Q_0 \sim 0.3$ but the power handling capability is only about 70 percent of its value at $Q_0 = 0$. If one does not, however, need the full power handling capability of the varactor one can probably find a varactor with lower power handling capability and larger f_c which, when driven at $Q_0 = 0$, will yield as high an efficiency. Due to space limitation only the $Q_0 = 0$ case is given. (For $Q_0 = 0.3$ see Grayzel.[13])

TABLE I

$\omega_c/\omega_3=$	10 000	100	50	25	10
$\overline{R}_1$	0.097	0.086	0.082	0.072	0.060
$\overline{R}_2$	0.096	0.097	0.098	0.107	0.139
$\overline{R}_3$	0.096	0.108	0.116	0.129	0.172
$\overline{X}_1$	0.179	0.176	0.176	0.175	0.178
$\overline{X}_2$	0.179	0.177	0.176	0.176	0.174
$\overline{X}_3$	0.179	0.184	0.188	0.193	0.203

TABLE II

$\omega_c/\omega_3=$	10 000	100	50	25	10
$\overline{R}_1$	0.145	0.135	0.129	0.120	0.095
$\overline{R}_2$	0.143	0.146	0.149	0.151	0.180
$\overline{R}_3$	0.143	0.152	0.156	0.170	0.211
$\overline{X}_1$	0.278	0.276	0.276	0.275	0.275
$\overline{X}_2$	0.278	0.277	0.276	0.275	0.279
$\overline{X}_3$	0.278	0.281	0.283	0.282	0.296

TABLE III

$\omega_c/\omega_3=$	10 000	100	50	25	10
$\overline{R}_1$	0.241	0.232	0.233	0.214	0.190
$\overline{R}_2$	0.238	0.247	0.246	0.253	0.276
$\overline{R}_3$	0.238	0.238	0.238	0.250	0.257

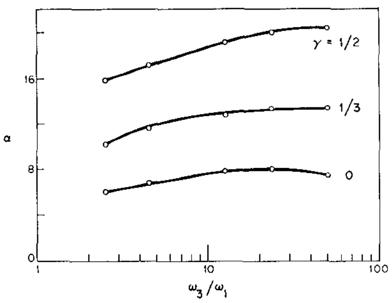

Fig. 4. α versus ω_3/ω_1 at maximum power point.

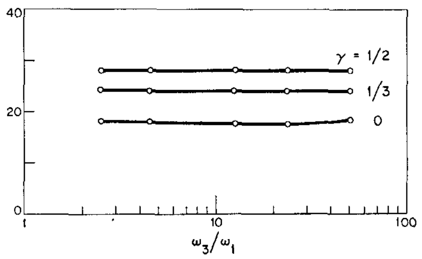

Fig. 5. β versus ω_3/ω_1 at maximum power point.

Impedances

Let us define normalized impedances by the relationships

$$R_K = \overline{R}_K S_{\max}/\omega_K$$
$$X_K = \overline{X}_K S_{\max}/\omega_K \quad (19)$$

The values of $\overline{R}_K$ and $\overline{X}_K$ do not vary appreciably with ω_3/ω_1. Values are given for different ratios of ω_c/ω_3 in Tables 1 through 3 for $\gamma=\frac{1}{2}$, $\frac{1}{3}$, and 0, respectively. For $\gamma=0$, $\overline{X}_K$ is equal to 0.500 and, hence, is not tabulated.

Bias Voltage

$V_0=0.111$ for $\gamma=\frac{1}{2}$, 0.139 for $\gamma=\frac{1}{3}$, and 0.184 for $\gamma=0$. The bias voltage does not vary with ω_c/ω_3 or ω_3/ω_1.

In Figs. 4 and 5 α and β are plotted VS ω_3/ω_1 for $\gamma=\frac{1}{2}$, $\frac{1}{3}$, and 0 at the maximum power point. For varactors with the same f_c $\gamma=0$ yields significantly higher efficiencies. For varactors with the same $V_B^2C_{\min}$ $\gamma=\frac{1}{2}$ gives the highest power handling capability.

Conclusion

The necessary design parameters have been presented to design a varactor USBUC. The impedances are given in Tables 1 through 3, the output power and efficiency in Figs. 1 through 3 and the bias voltage is given above.

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