Microwave Varactor Tuned Transistor Oscillator Design

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Abstract—**An analysis is made of the common base microwave transistor oscillator circuit which uses a varactor in series with the collector to tune over octave bandwidths. Equations are derived giving the required feedback capacitances and resonating elements required for octave tuning. Normally, the collector-emitter capacitance C_{ce} is made approximately equal to the transistor collector capacitance C_c. The emitter-base capacitance C_{eb} is important only at very high frequencies. It is shown that a high-Q varactor must be used and that only a limited amount of collector-base capacitance C_{cb} may be added if the circuit is to be resonated over an octave. The output power for such a circuit is normally about 1/5 the maximum power available from the transistor.**

Experimental oscillators were made from 0.5 to 1 GHz and 1 to 2 GHz which substantially verified the analysis. Using the TIXS13 transistor, an output power of 200 mW was obtained from 430 to 860 MHz tuning from −2 to −115 volts. In the 1 to 2 GHz range a TIXS13 transistor oscillator was tuned from 1.09–1.96 GHz with about 40 mW power tuning from −2 to −115 volts. By use of a lower case capacitance varactor, the 1 to 2 GHz oscillator could be made to tune over the full octave.

I. Introduction

With the development of microwave transistors, it is now possible to develop solid-state varactor or YIG tuned microwave transistor oscillators with octave bandwidths. Such voltage tunable oscillators are already beginning to replace backward wave oscillators and klystrons for low-frequency application. They possess the advantages of small size and weight, long life, and greater reliability. With the development of S-band transistors coupled with the use of collector base harmonic generation [1]–[3] wideband varactor tuned transistor oscillators could be constructed up to X-band frequencies, especially for local oscillator applications.

This paper presents an analysis of a varactor tuned transistor oscillator giving conditions for oscillation, theoretical tuning curve, and load matching conditions for optimum performance. Experimental results from a 0.5 to 1 GHz and a 1 to 2 GHz varactor tuned oscillator are given. Other design techniques such as the use of collector base multiplication are also considered.

While YIG tuning is another way of oscillator tuning, it is not considered here. YIG possesses the advantages of higher Q and lower tuning voltage and a linear voltage vs. frequency tuning curve. Varactors possess the advantage of utility at low frequencies, adaptability to monolithic microwave circuitry, and low current requirement for tuning. This paper only considers varactor tuning although YIG tuning theory can be extended from this paper.

II. Theoretical Analysis

The first thing to be determined is an adequate simplified high-frequency equivalent circuit for the transistor. Second, a circuit must be chosen which will accomplish octave tuning. Third, an analysis must be made to determine the value of the circuit elements which will satisfy the oscillation criteria.

A. *Simplified High-Frequency Equivalent Circuit*

In Fig. 1(a) a high-frequency transistor equivalent circuit is shown [4], [5]. Here we define

r_e = emitter resistance
r_b = base resistance
C_e = emitter capacitance
μ = reverse transfer ratio
C_c = collector capacitance
α = common base forward current transfer ratio
I_e = emitter current
V_c = voltage across collector capacitance
C_{eb} = emitter base capacitance
C_{cb} = collector base capacitance
C_{ce} = collector emitter capacitance.

Here C_{eb}, C_{cb}, C_{ce} include the transistor package capacitances plus any externally added capacitance. The current transfer ratio or current gain α of the transistor as a function of frequency is derived from transistor theory. The expression is, however, a complex equation involving transcendental functions. An approximate simplified form is used for representation with lumped networks and is

$$\alpha(\omega) = \frac{\alpha_0' e^{j\omega T}}{1 + j\omega/\omega_\alpha} \qquad (1)$$

where α_0' is the low-frequency value of the ratio of the collector current I_c to the emitter current I_e. The function $e^{j\omega T}$ results from the transit time of carriers across the base. The α frequency dependence of (1) can be further simplified for $\omega T \ll 1$ which is the case for the

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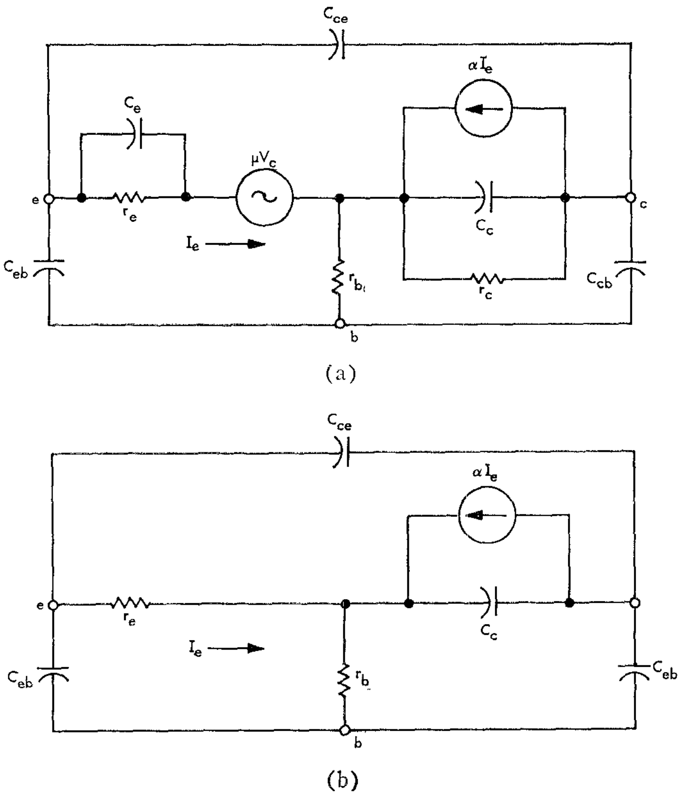


Fig. 1. Transistor equivalent circuit. (a) Basic equivalent circuit. (b) Simplified equivalent circuit.

transistor we used. Then, (1) can be written

$$\alpha(\omega) = \frac{\alpha_0}{(1 + j\omega/\omega_\tau)} \tag{2}$$

where

$$\omega_\tau = \omega_\alpha(1 - \omega^2 T/\omega_\alpha)/(1 + \omega_\alpha T)$$

and

$$\alpha_0 = \alpha_0'/(1 - \omega^2 T/\omega_\alpha).$$

The expression of (2) is the one we shall use in our analysis. In this case it will be assumed that $\omega^2 T/\omega_\alpha \ll 1$ so that α_0 is a constant. In addition, ω_τ will be assumed to include the reduction in α due to C_e. This latter effect is usually quite small anyway, since at high emitter currents r_e is extremely small and effectively shorts out C_e. That is, r_e is given approximately as $r_e = kT/qI_e$. So at 100 mA r_e is only $\frac{1}{4}$ of an ohm and therefore C_e can be neglected. Also, $r_e \ll r_b$ since r_b is usually a couple of ohms. The resistance r_c is always negligible at microwave frequencies.

One further simplification can be made, this is to neglect μV_c. The effect of this generator is nearly always negligible at high frequencies since $\mu \ll 1$ and $V_c \propto I_e/\omega C_c$, so V_c becomes very small at high frequencies. With these simplifications, the resulting equivalent circuit is shown in Fig. 1(b). The next thing to be determined is the circuit to be used.

B. Circuit Choice

There are a number of possible circuits that could be used. The circuit choice was made after an experimental investigation of possible circuit configurations. There were two primary constraints which limited the oscillator circuit choice. First, the transistor and its package and, second, the large capacitance change required to tune the oscillator over an octave.

Because of the large transistor package capacitances, parallel tuning of the transistor oscillator could not be done at microwaves since too large a varactor junction capacitance would be required to tune over an octave. While high capacitance varactors are available, their Q is too low to make them usable. This means that a series tuning circuit must be employed with additional shunt capacitance added across the varactor and resonating inductance to assure that resonance occurs in series with the varactor and that tuning over an octave can be accomplished.

The varactor tuning and resonating equivalent circuit is shown in Fig. 2(a). Here, C_0 is the junction capacitance, R_s' is the series resistance, C_s' is the varactor case capacitance, L_s is the series tuning inductance, and C_p is the total shunt capacitance across the varactor required to tune over an octave. C_p would also include any transistor package capacitance that the varactor is placed across.

For purposes of analysis this equivalent circuit can be further simplified as shown in Fig. 2(b). Here R_s and C_s now represent the effective resistance and capacitance corresponding to R_s' and C_s' if the case capacitance were only across the junction capacitance rather than both R_s and C_0. Over the entire tuning range this circuit at its terminals must look only like a net inductance and resistance. Therefore, the circuit of Fig. 2(b) will be replaced by that of Fig. 2(c) when solving for the oscillation condition.

The circuit of Fig. 2(b) showing the resonating and tuning elements provided for the transistor oscillator circuit could, by the way, also represent the equivalent circuit for a YIG sphere. If no additional inductance is assumed added to the transistor, this circuit represents the complete resonating element for the oscillator. Three possible circuit configurations could be used, forgetting for the time being, the load. The varactor circuit could be placed across C_{eb}, or C_{ce}, or C_{cb}. All three of these circuits were constructed and the varactor tuned over the range. The circuit used no external load but the field was probed lightly to determine if it was oscillating. Additional capacitances as needed were added to sustain oscillation. The varactor series resistance actually acted as a load.

It was found that only with the varactor circuit tank across C_{cb} could large oscillation be sustained over the entire capacitance tuning range. This indicated that oscillation criteria could not adequately be met in the

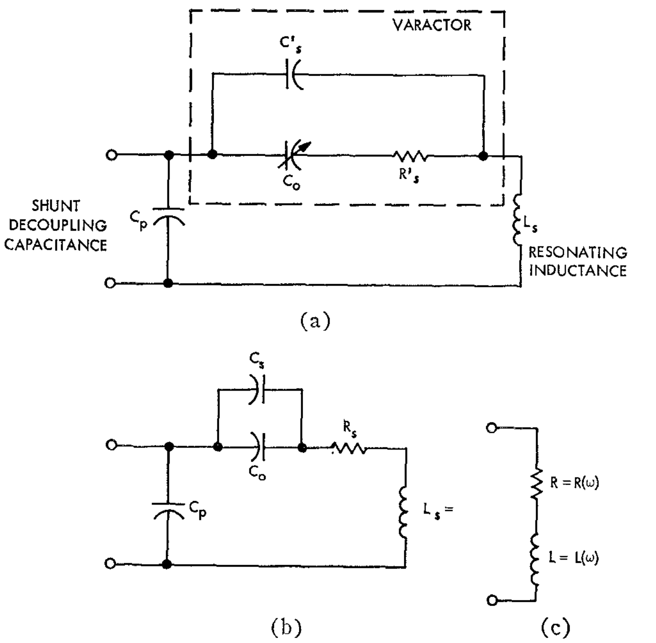


Fig. 2. Varactor tuning circuit. (a) Complete equivalent circuit of varactor, resonating inductance and shunt decoupling. (b) Simplified circuit of (a). (c) Reduced circuit of (b) for purposes of analysis.

other configurations over the entire tuning range. Analysis done on the other possible circuit configurations verified this conclusion.

The transistor package used (the TIXS13) was suited to a common base configuration with the tank circuit across the collector base terminals, thus, this circuit configuration was used [6].

C. Circuit Analysis

1) Oscillation Conditions [7]: The oscillator circuit including the varactor and load is shown in Fig. 3. Here, using the Thevenin equivalent for αI_e and C_c, we define

$$\frac{\alpha}{j\omega C_c} = -R_\alpha\left(1 + j\frac{\omega_\tau}{\omega}\right) \tag{3}$$

where

$$R_\alpha = \frac{\alpha_0 \dfrac{\omega}{\omega_\tau}}{\left(1 + \dfrac{\omega^2}{\omega_\tau{}^2}\right)\omega C_c}\,. \tag{4}$$

The matrix for this circuit is then given as follows

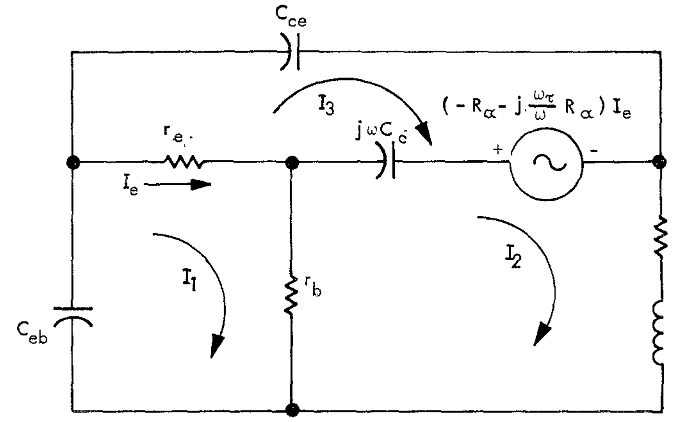


Fig. 3. Transistor oscillator equivalent circuit.

where

$$C_t = \frac{C_c C_{ce}}{C_c + C_{ce}}\,.$$

The conditions for oscillation are that the real and imaginary parts of this matrix be equal to zero [8]. Before computing the real and imaginary parts, the matrix can be simplified by neglecting the small terms. That is, $r_e \ll r_b \ll R_\alpha$, $r_b \ll \omega_\tau/\omega^2 C_{ce}$, and $r_b \ll 1/\omega C_{eb}$ in the high-power high-frequency transistor used. We may further assume we have a load circuit Q, greater than 3, over most of the tuning range. In this case further simplification can be made since this means that $\omega L \gg (r_b + R)$. With these assumptions the circuit matrix can be computed as follows.

Imaginary Part

$$-\frac{1}{\omega C_{eb}}\left\{\omega L\left(\frac{1}{\omega C_t} - R_\alpha\frac{\omega_\tau}{\omega}\right) - \frac{1}{\omega^2 C_c C_{ce}}\right\} + r_b R_\alpha \omega L + \frac{r_b R_\alpha}{\omega C_c} + \frac{r_b R_\alpha}{\omega C_t} = 0. \tag{6}$$

We observe that since $1/\omega C_{eb} \gg r_b$ then terms in r_b might be neglected. However, to see the effect of ωC_{eb} should it be made large, these terms will be included. When terms in r_b are neglected, then

$$\omega L = \frac{1}{\omega^2 C_c C_{ce}} \frac{1}{\left(\dfrac{1}{\omega C_t} - R_\alpha\dfrac{\omega_\tau}{\omega}\right)}\,. \tag{7}$$

$$\Delta = \begin{vmatrix} r_e + r_b - j\left(\dfrac{1}{\omega C_{eb}}\right) & -r_b & -r_e \\ -r_b + R_\alpha + j\dfrac{\omega_\tau}{\omega}R_\alpha & r_b + R + j\left(\omega L - \dfrac{1}{\omega C_c}\right) & -R_\alpha - j\left(\dfrac{\omega_\tau}{\omega}R_\alpha - \dfrac{1}{\omega C_c}\right) \\ -r_e - R_\alpha - j\dfrac{\omega_\tau}{\omega}R_\alpha & j\left(\dfrac{1}{\omega C_c}\right) & r_e + R_\alpha + j\left(\dfrac{\omega_\tau}{\omega}R_\alpha - \dfrac{1}{\omega C_t}\right) \end{vmatrix} \tag{5}$$

Usually

$$\frac{1}{\omega C_t} \gg R_\alpha \frac{\omega_T}{\omega} \text{ at high frequencies}$$

so

$$\omega L = \frac{1}{\omega(C_c + C_{ce})}. \tag{8}$$

Resubstituting this in the small term in r_b in (6) we obtain

$$\omega L\left(\frac{1}{\omega C_t} - R_\alpha \frac{\omega_T}{\omega}\right) - \frac{1}{\omega^2 C_c C_{ce}} = \frac{C_{eb} r_b R_\alpha}{C_{t2}} \tag{9}$$

where

$$\frac{1}{C_{t2}} = \frac{1}{C_c} + \frac{1}{C_t} + \frac{1}{C_c + C_{ce}}. \tag{10}$$

Substituting (9) into the matrix (5) and making the previous approximations, the real part can be computed as follows.

Real Part

$$R_\alpha\left[\frac{1}{\omega^2 C_c C_{ce}\left(\frac{1}{\omega C_t} - R_\alpha \frac{\omega_T}{\omega}\right)} + \frac{C_{cb} f_T r_b}{C_{ce} f} - \frac{r_b^2 \omega C_{eb}^2}{C_{t2}}\right] = (r_b + R)\left(\frac{1}{\omega C_t} - R_\alpha \frac{\omega_T}{\omega}\right). \tag{11}$$

Rewriting, substituting (4) for R_α and neglecting $R_\alpha \omega_T/\omega \ll (1/\omega C_t)$ (since $C_t \gg C_c$), we have

$$1 + \frac{\omega^2}{\omega_T^2} = \frac{\alpha_0 C_{ce}}{\omega_T (C_c + C_{ce})^2 (r_b + R)} \cdot \left[1 + \omega_T r_b C_{eb}\left(1 + \frac{C_c}{C_{ce}}\right)\left(1 - \frac{\omega_T C_{eb} r_b C_{ce}}{C_{t2}}\right)\right]. \tag{12}$$

From this the value of C_{ce} and C_{eb} giving the maximum frequency of oscillation can be computed. This is

$$C_{ce} = C_c; \frac{\omega_T C_{eb} r_b C_{ce}}{C_{t2}} \doteq \frac{1}{2}. \tag{13}$$

As expected, the term in $\omega_T r_b C_{eb}$ only adds about 20 percent to the value of the right-hand side of the equal sign. However, if C_{ce} is somewhat less than optimum, this can be a more significant term. In practice it is found to be important at frequencies above ω_T.

2) Varactor Tuning Circuit: Let us again turn to the varactor tuning circuit of Fig. 2(b) to compute the effective resistance R and the resonance condition. The result is, assuming resonance and (6) to hold,

$$R = R_s\left[\left(1 - \frac{2\omega C_p + \omega(C_c + C_{ce}) \pm \omega(C_c + C_{ce})\sqrt{1 - 4\omega^2 C_p^2 R_s^2 (1 + C_p/(C_c + C_{ce}))^2}}{2\omega(C_p + C_{ce} + C_c)}\right)^2 + \omega^2 C_p^2 R_s^2\right]^{-1}. \tag{14}$$

Resonance Condition

$$\omega L = \frac{1}{\omega(C_0 + C_s)} + \frac{2\omega C_p + \omega(C_c + C_{ce}) \pm \omega(C_c + C_{ce})\sqrt{1 - 4\omega^2 C_p^2 R_s^2 (1 + C_p/(C_c + C_{ce}))^2}}{2\omega C_p(\omega C_p + \omega C_c + \omega C_{ce})}. \tag{15}$$

To have a series resonance, $4\omega^2 C_p^2 R_s^2 (1 + C_p/(C_c + C_{ce}))^2 < 1$, and a minus sign must be used. This is, then, a condition for proper varactor loading. Notice that R_s may also include a series load resistance term. If $2\omega C_p R_s$ becomes greater than unity a reactive term appears in the oscillator which quickly causes the oscillator to stop oscillating. (The term C_p is essentially equal to the total collector base case plus external capacitance from collector to base.)

Notice, from (14), that if the term $4\omega^2 C_p^2 R_s^2 (1 + C_p/(C_c + C_{ce}))^2$ is not less than unity, so that the oscillator will not oscillate, it can be reduced by increasing C_{ce} beyond the optimum value given in (11). This is often the case and often determines the value of C_{ce} rather than (11).

Under the condition $4\omega^2 C_p^2 R_s^2 (1 + C_p/(C_c + C_{ce}))^2 \ll 1$ the series resonance condition reduces to

$$\omega L = \frac{1}{\omega(C_0 + C_s)} + \frac{1}{\omega(C_p + C_c + C_{ce})}. \tag{16}$$

The normal capacitance voltage characteristic is given as

$$C(V) = \frac{C(V = 0)}{\left(1 - \frac{V}{\phi}\right)^n} \tag{17}$$

where n typically is 0.46 and ϕ is 0.7 for a silicon diode. V is the bias voltage and $C(V=0)$ is the zero bias capacitance. Substituting typical capacitance values in (16) and (17), a tuning curve can be drawn as shown in Fig. 4. Notice that it does not have a straight line log characteristic due to the parasitic capacitance C_s at the high-frequency end and due to $C_s + C_c + C_p$ at the low-frequency end.

3) Load Circuit: A number of possible load circuits were tried experimentally and the best place to insert the load was in series with the varactor. This seems a reasonable place since the varactor resistance appears in series with r_b of the transistor and remains constant over the range.

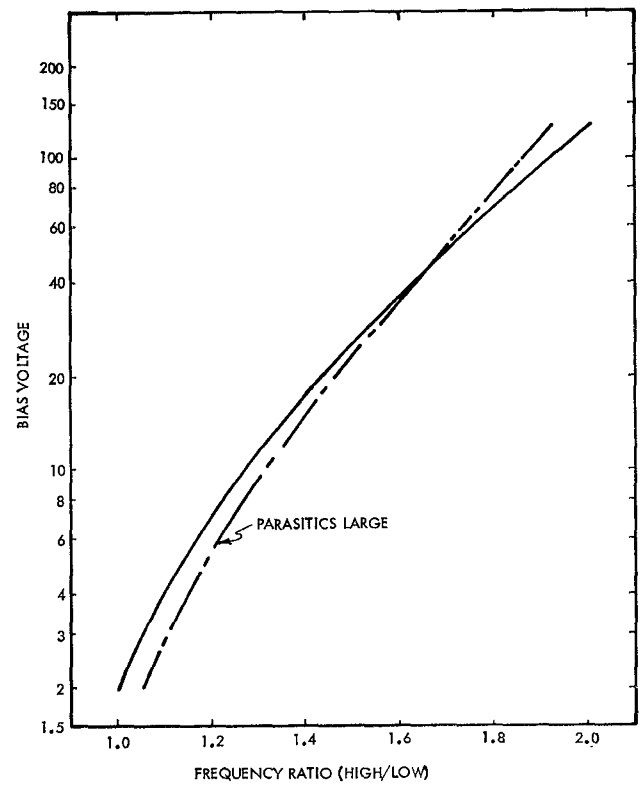
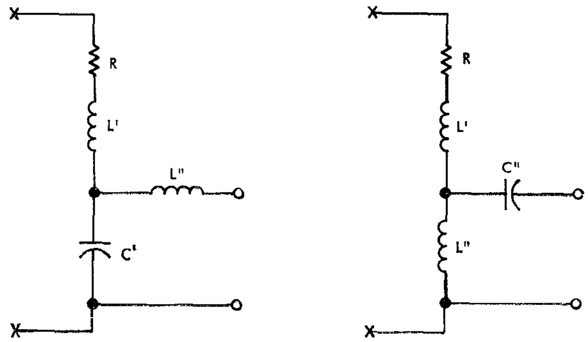


Fig. 4. Theoretical transistor oscillator, varactor tuning curve.



(a) (b)

Fig. 5. Load coupling circuits.

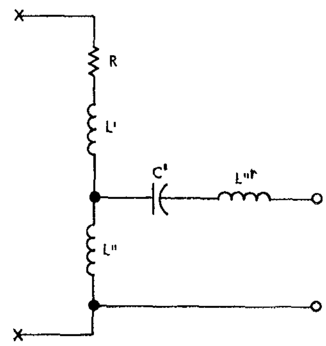


Fig. 6. Broadband load coupling circuit.

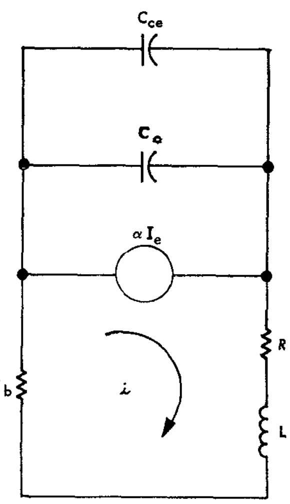


Fig. 7. Transistor oscillator equivalent circuit for computing power out.

The series loading normally will be of the same order of magnitude as the varactor resistance. This means some impedance transformation must be used. This can be accomplished by the usual impedance transformation techniques as shown in Fig. 5. The first technique shown in Fig. 5(a) involves coupling into the series line across a capacitor. This unfortunately tends to reduce the tuning range of the varactor circuit. The second technique involves coupling into the series line across an inductance. This technique is preferable since it adds no series capacitance. This can be easily done by tapping on to the series inductance of the varactor tuning circuit. Notice here that a series capacitance is also required. The capacitance is necessary to provide a flatter amplitude vs. frequency response. In the next section, where we consider the power out, it will be seen that a load is required which increases with frequency. The inductive tap and series capacitance circuit provide just this. Additional response improvement can be obtained by use of an inductance in series with the output capacitance. This tends to provide a broader bandwidth output circuit in a fashion similar to an m derived filter. This circuit is shown in Fig. 6. Due to transistor parameter variations over the tuning range experimental compensation techniques must be used to some extent.

4) Power Out: The power out can be calculated from the circuit of Fig. 3, assuming that it is oscillating such that the nonlinearity of the transistor provides the limiting voltage and current so that a full emitter current flows at the RF frequency. The circuit of Fig. 3 may be simplified to that shown in Fig. 7. Here the emitter base circuit is neglected including neglecting r_e. From this circuit the power P_R into the effective resistance R can be computed. This is

$$P_R = \frac{\alpha_0^2 I_e^2 R}{\omega^2 (C_c + C_{ce})(r_b + R)^2 (1 + \omega^2/\omega_r^2)}. \tag{18}$$

Using (14) and simplifying we have

$$R = (R_s + R_L)(1 + C_p/(C_c + C_{ce}))^2 \tag{19}$$

where now a load resistance is assumed in series with

$$R_s \text{ and } 4\omega^2 C_p^2 (R_s + R_L)^2 \left(1 + \frac{C_p}{C_c + C_{ce}}\right)^2 \ll 1.$$

The resultant power out into the load resistance P_L becomes

$$P_L = \frac{\alpha_0^2 I_e^2 R_L (1 + C_p/(C_c + C_{ce}))^2}{\omega^2 (C_c + C_{ce})^2 [r_b + (R_s + R_L)(1 + C_p/(C_c + C_{ce}))^2]^2 (1 + \omega^2/\omega_T^2)}. \tag{20}$$

This can be compared with the maximum available output power P_{AV} which is

$$(R_s = 0, C_p = C_{cb}, C_{ce} = C_c, R_L(1 + C_p/(C_c + C_{ce}))^2 = r_b) \quad P_{AV} = \frac{\alpha_0^2 I_e^2}{4\omega^2 C_c^2 r_b (1 + \omega^2/\omega_T^2)}. \tag{21}$$

This equation is valid only at high frequencies. Computing the ratio

$$\frac{P_L}{P_{AV}} = \frac{16 r_b R_L (1 + C_p/(C_c + C_{ce}))^2}{(1 + C_{ce}/C_c)^2 [r_b + (R_L + R_s)(1 + C_p/(C_c + C_{ce}))^2]^2 \left(1 + \frac{C_{cb}}{2C_c}\right)^2}. \tag{22}$$

Usually, for maximum power transfer

$$r_b = R_L (1 + C_p/(C_c + C_{ce}))^2.$$

Thus

$$\frac{P_L}{P_{AV}} = \frac{4}{\left(1 + \frac{C_{ce}}{C_c}\right)^2 \left(1 + \frac{R_s}{2R_L}\right)^2 \left(1 + \frac{C_{cb}}{2C_c}\right)^2}. \tag{23}$$

Recall that for proper oscillation condition $C_{ce} = C_c$, a condition usually met at microwaves. Then

$$\frac{P_L}{P_{AV}} = \frac{1}{\left(1 + \frac{R_s}{2R_L}\right)^2 \left(1 + \frac{C_{cb}}{2C_c}\right)^2};$$

typically $C_{cb} \approx C_c$. (24)

Thus, a transistor rated for 1 watt power output will have a nominal output in this circuit of approximately 200 mW if $R_L = R_s$. This has been found to be true in practice. Some improvement over this is realized if C_{eb} is used.

One important factor observed from (18) is that the power out for a fixed load falls with frequency. This means that a load circuit should be used which effectively causes R_L to increase with frequency. This can be accomplished and, in fact, is inherent in inductive tapping of the coil since the effective series load resistance is $R_{Leff} = \omega^2 L^2 / R_L$. Thus, simply tapping into the inductor at the correct place should provide a fairly uniform amplitude characteristic. This has been found to be true experimentally although usually some series capacitance is still required to achieve a flat response.

III. Experimental Results

A. Design Approach

Initial calculations based on the preceding analysis revealed that there was sufficient collector base capacitance in the transistor case capacitance, so that no additional collector base capacitance was required for $C_c = C_{ce}$.

According to the TIXS13 transistor specification, the common base output capacitance is about 8pF $= C_{ce} + C_c$. Adding to this C_{cb}, the total output capacitance is about 12pF. Using an AEL varactor with 5pF capacitance at -4 volts, and a package capacitance of 0.20pF, the maximum to minimum capacitance swing from 120 volts to -2 volts is 6.5–1.1pF or a factor of 5.9 to 1. Including case capacitance this becomes a factor of 5.2 to 1 or 6.7–1.3pF. Considering finally the output capacitance [see (13)] this reduces to 3.65 to 1. An additional shunt capacitance of about 5pF must be added to make this greater than a 4 to 1 change in capacitance.

Referring to (14) the loading condition calculation is carried out at 2 GHz where $C_p = (5+4)$pF $= 9$pF, $C_{ce} = 4$pF, $C_c = 4$pF and R_s is about 1.2 ohms. The factor $2\omega C_p R_s (1 + C_p/C_{ce}) = 0.58$. This is less than unity and therefore satisfies this load condition. It is, however, close enough to unity so that at slightly higher frequencies this condition is not satisfied and the oscillator ceases to oscillate. This condition was observed in the 1 to 2 GHz oscillator.

These values also allow us to determine how large R_s is in the collector circuit. This is determined from (19). The result is $R = 4.5 R_s$. It is important to have C_p as small as possible to minimize this effect.

The circuit configuration for the 1 to 2 GHz oscillator is shown in Fig. 8. The varactor load circuit consisted of a coaxial cavity. The resonating inductive length was less than $\lambda/10$ and therefore approximates a lumped inductance. The output was tapped onto the center conductor as shown with a series inductance and capacitance to broadband the circuit. Conditions for oscilla-

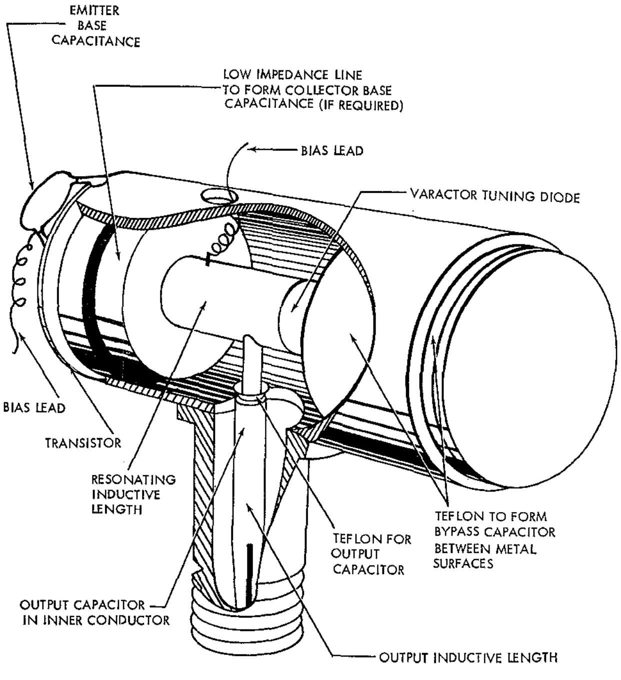


Fig. 8. 1 to 2 GHz transistor oscillator construction (bias omitted).

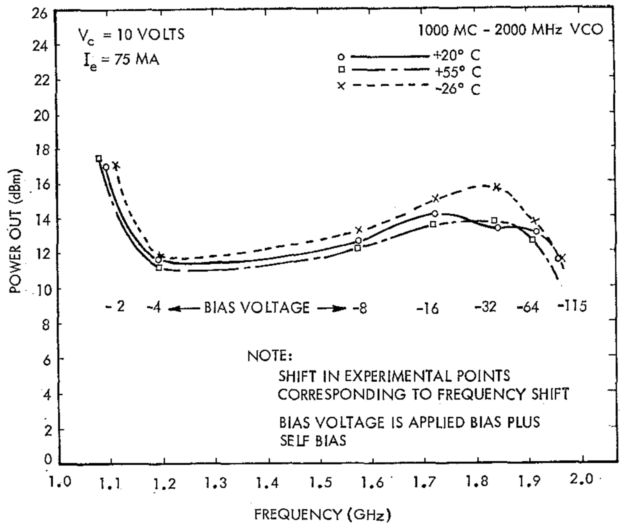


Fig. 9. 1 to 2 GHz transistor oscillator power out vs. frequency.

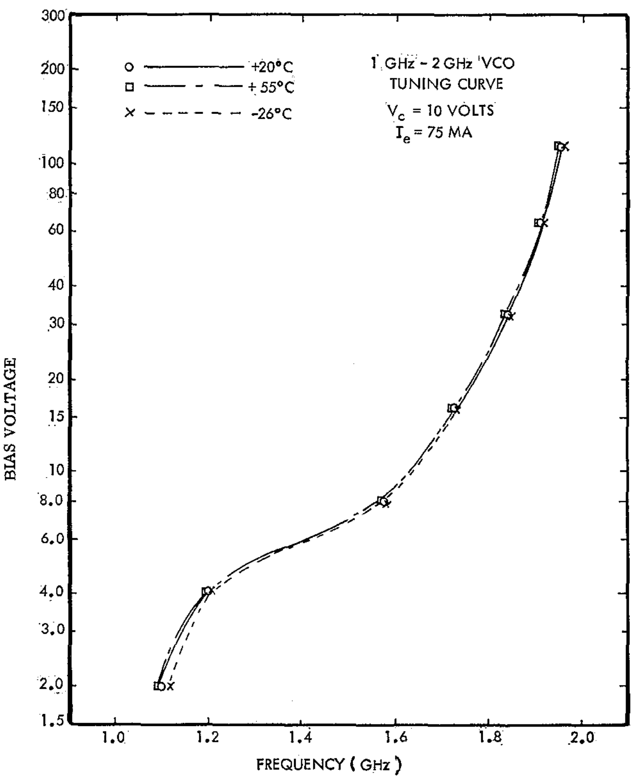


Fig. 10. 1 to 2 GHz transistor oscillator, varactor tuning curve —frequency vs. bias voltage.

tion at 2 GHz were quite critical. It was found that by using a small capacitor from emitter to base that this critical condition was greatly alleviated.

A typical response curve for this oscillator is shown in Fig. 9. The curve shown does not quite cover an octave since in this oscillator a varactor was used which has a case capacitance of 0.28pF rather than 0.2. Later design improved on this bandwidth. Notice from the figure that the temperature does not too greatly affect the performance. Notice, too, the rapid fall off at frequencies approaching 2 GHz. The tuning curve for this circuit is shown in Fig. 10. Here a reverse S-shape can be seen which is due to the varactor drawing some current in the forward direction.

An oscillator for UHF was also constructed as shown in Fig. 11. Here the varactor tuning circuit was simply a short wire loop. The entire circuit had to be kept in a shielded can. No external capacitances were required, in this case, to tune over the range. It should be added that actually two other factors enter to increase the tuning range if properly done. One is the load circuit tapping into the inductance can increase the tuning range slightly. The second factor is seen in (7) and (8) where we had neglected $R_a\omega_\tau/\omega$. At the lower frequencies this term becomes more important and tends to enhance the tuning range. This is actually observed in the UHF oscillator as seen in the response curve of Fig. 12. While the same varactor was used as for the 1 to 2 GHz case this oscillator has a wider bandwidth, in fact greater than 2 to 1. However, temperature has a stronger reaction to this effect.

The tuning curve for the UHF oscillator is shown in Fig. 13. Here the curve at the higher end shows the slight reverse S-shape and at the low end it shows a slight forward S-shape.

Another frequency tuning effect also should be mentioned. At low bias voltages the varactor begins to conduct, adding rectified bias voltage to the actual bias. Typically this voltage was 2.0 volts. At the same time, the nonlinear capacitance of the characteristic tends to raise the capacitance to a larger value. This makes it very difficult to determine the actual capacitance at zero bias or -2 volts.

It should be mentioned here that since there is a rectified voltage developed across the varactor at low

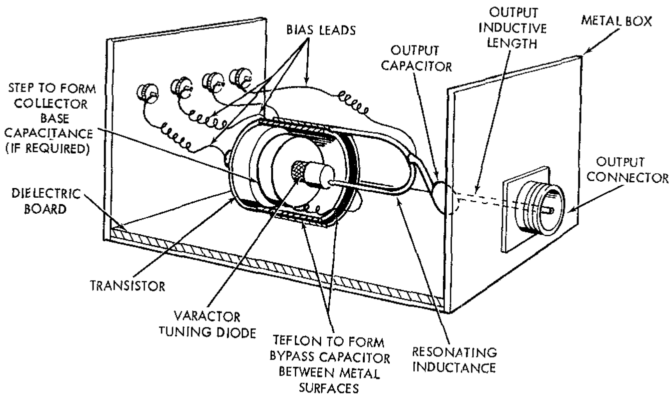


Fig. 11. 500 to 1000 MHz transistor oscillator construction (bias omitted).

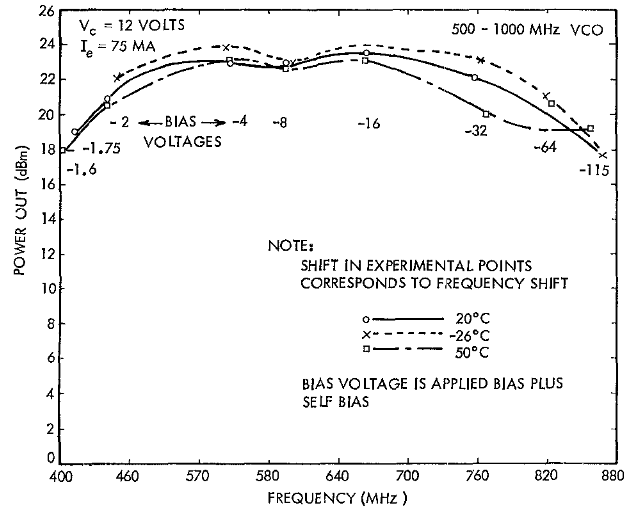


Fig. 12. 500 to 1000 MHz transistor oscillator power out vs. frequency.

bias voltages, it is necessary to have the diode capacitance large to minimize this voltage. This voltage is approximately $I_c/\omega(C_c+C_{ce})$. This consideration is not too important at microwaves where ω is large but can be important at low UHF frequencies.

IV. Other Design Techniques

While the techniques to be described have not been tried experimentally they do show some of the oscillator improvements that are possible.

A. Multiple Series Tuning Varactors

This technique was suggested by Irvin et al. [9] for harmonic generators and worked quite well. Essentially this technique, that of using more than one tuning varactor in series, allows one to use lower breakdown and therefore higher Q and, if desired, higher capacitance varactors. The price to be paid is that a larger tuning voltage must be used.

As an example using two diodes, the capacitance is one-half that of one and the resistance is twice that of one. Now, however, no additional capacitance is required in C_p to tune over the 2 to 1 frequency range.

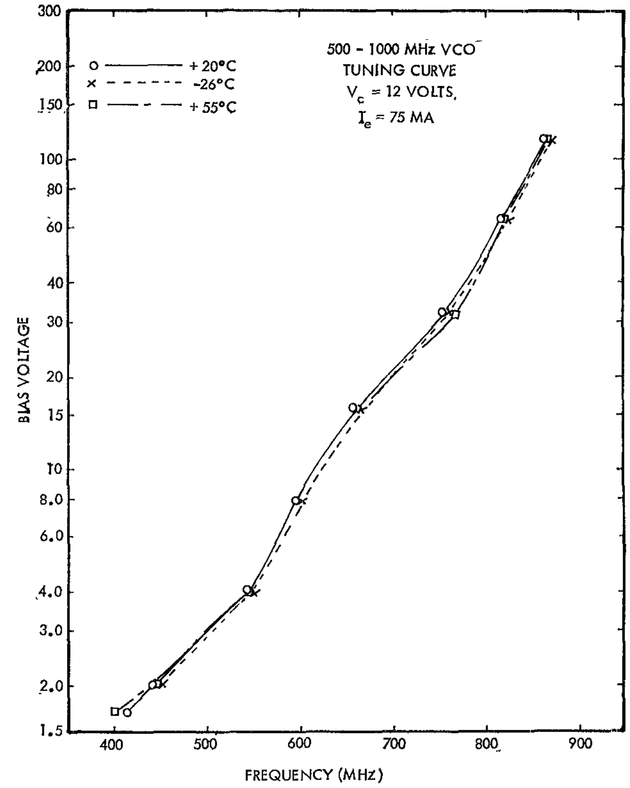


Fig. 13. 500 to 1000 MHz transistor oscillator, varactor tuning curve frequency vs. bias voltage.

Thus, $C_p = 4$pF and therefore the load factor is 0.36 and the effective resistance R is 2.25 R_s. This is quite an improvement especially at 2 GHz frequencies. Going to 4 varactors with slightly more than half the resistance and twice the capacitance gives us the same improvement, except now the parasitic shunt capacitance has less effect and the tuning range is further extended.

B. Use of Amplifiers

Since the varactor tuned transistor oscillator has a reduced efficiency when compared with a nonlossy oscillator, a possible way of improving the oscillator is to use a transistor amplifier stage following the oscillator as shown in Fig. 14. This of course only works up to f_T since about this frequency insufficient broadband gain can be achieved to justify this technique.

C. Collector Base Multiplication

To obtain tuning at higher frequencies, a possible approach is to use the collector base diode as a varactor doubler. In this case, the collector base circuit is tuned to both the fundamental and the second harmonic. Such a circuit is shown in Fig. 15. This circuit has been observed to double at low UHF frequencies but has not been tested at S-band. This type of operation has been considered at a single frequency and should operate using varactors to tune over a wide range. An alternate approach if a common collector oscillator is used would be to tune the collector base circuit to the fundamental and the emitter base circuit to the desired harmonic.

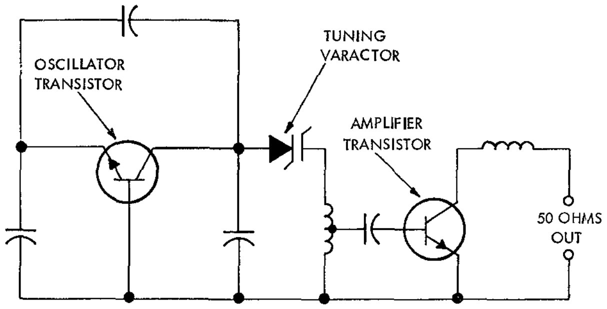


Fig. 14. Oscillator amplifier combination (excluding bias circuitry).

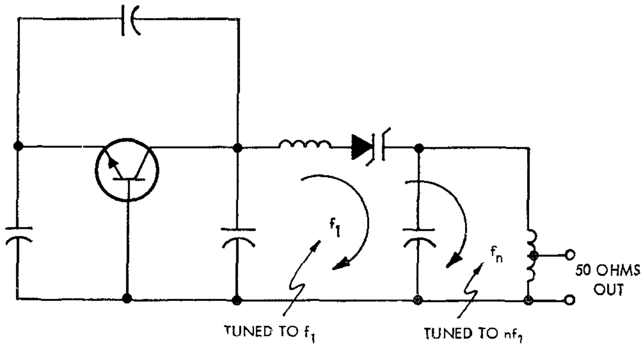


Fig. 15. Transistor oscillator doublers combination (excluding bias circuitry).

V. Conclusions

A common base varactor tuned transistor amplifier circuit has been analyzed and the circuit conditions for proper oscillation have been defined. Two such oscillators were constructed, one at *L*-band and the other at UHF. Both performed as expected, the results conforming quite well to what the analysis predicted. It is expected that the use of multiple diodes can improve the performance of the circuit. Collector base multiplication can be used to extend this operation into *S*-band.

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Propagation in a Microwave Model Waveguide of Variable Surface Impedance—Theory and Experiment

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***Abstract*—In this paper propagation in a model terrestrial waveguide is investigated. The surface impedance of the waveguide boundary is assumed to vary along the path of propagation. A quasi-optical approach is used to derive the solution for the case of an abrupt variation in the surface impedance. The reciprocity theorem is employed to facilitate that solution for both directions of propagation. Experimental verification of this technique is obtained from measurements in the model waveguide.**

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I. Introduction

IN EARLIER model studies[1,2,3] of radio propagation in a nonuniform terrestrial waveguide, the variations of the electrical properties of the ionosphere

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