Design of Efficient Broadband Varactor Upconverters

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This paper describes a design procedure for optimizing the performance of a varactor upconverter microwave power amplifier with respect to maximum pump efficiency; the procedure gives explicit diode and circuit parameters and operating levels. An optimum diode is selected by inclusion of an empirical relation between diode breakdown voltage and cutoff frequency. A fully driven abrupt junction diode is assumed.

The bandwidth of an upconverter is limited by the intermediate frequency input circuit which typically has a 3 dB bandwidth of about 10 percent. We describe a method of obtaining broadband operation where the interface between the intermediate frequency source and the varactor diode is mismatched in an optimum way. Analysis shows that the frequency variation in mismatch loss just compensates for the frequency variation in upconverter gain predicted by the Rowe–Manley relations.

The design procedure is illustrated by a 300 MHz to 10.960 GHz varactor upconverter built for use in the transmitter of the short hop radio system experiment. A bandwidth of 120 MHz between 1.4 dB points represents a bandwidth of more than 40 percent at the intermediate frequency. An output power of +16 dBm was obtained using a pump power of +20 dBm giving a pump efficiency of 40 percent. The normal input to the driver amplifier is +3 dBm giving an overall gain of +13 dB. The upconverter operates over a temperature range of −40°F to +140°F with only a small change in bandshape and output power.

I. INTRODUCTION

In many microwave radio relay systems the transmitted microwave signal is obtained by upconverting the IF signal in a mixer or modulator and amplifying it in a microwave power amplifier, such as a traveling wave tube, to a power level set by system requirements. By use of a varactor upconverter as a microwave power amplifier the functions of upconversion and amplification can be performed by one device. In fact, for all-solid-state radio systems the varactor upconverter is the only broadband microwave power amplifier available. Therefore, in radio systems such as those described in companion papers, broadband, efficient, rugged varactor upconverters are required as power amplifiers.[1,2]

Although several analyses of varactor upconverters have been published, design has required a trial and error procedure with its inherently uncertain results. In this paper we describe a design procedure for optimizing the performance of an upper sideband varactor upconverter with respect to maximum pump efficiency. This procedure gives diode and circuit parameters explicitly in terms of operating power levels. We also describe a method of obtaining broadband operation. Although the procedure is derived using specific frequencies and the maximum efficiency optimization, the basic method can be used for any frequency and optimization.

The design procedure and broadbanding method are illustrated by the design of an upconverter for the experimental radio system described by Ruthroff and others.[3] The system requires a bandwidth of 120 MHz between 1 dB points and has an intermediate frequency of 300 MHz. A maximum of 100 mW (+20 dBm) of pump power was available; the available IF power was +3 dBm. The objective was maximum output power consistent with the foregoing constraints and the limitations imposed by the environment. The performance of the two upconverters required in the experiment was essentially the same, so the discussion is primarily of the 300 MHz to 10.960 GHz upconverter.

II. SUMMARY

The limited pump power available at 11 GHz requires maximum pump efficiency upconverter operation. From the large signal analysis of Penfield and Rafuse and an empirical relation between diode cutoff frequency and breakdown voltage, the output power versus pump power for fully driven maximum efficiency operation is found.[4–6] Then, given minimum output power and maximum pump power available, an operating range and all diode circuit parameters and diode characteristics are specified. The details of this procedure are given in Section III.

The upconverter is a straight-through type with a single diode (see

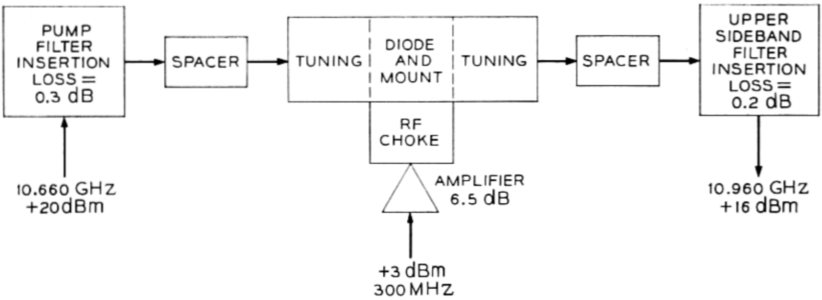

Fig. 1 — Block diagram of upconverter.

the block diagram in Fig. 1). The RF circuit uses a standard WR90 X-band waveguide and consists of properly located signal and pump filters for providing the required terminations, tuning sections for matching input and output signals, and the diode mount and impedance transforming capacitor. The diode-to-waveguide impedance transformation circuit consists of the diode inductance and an adjustable shunt capacitance formed by an adjustable sleeve concentric with the diode. This method of matching eliminates the need for reduced height waveguide and associated transitions and allows the filters to be close to the diode with a corresponding reduction of band-shape ripples. The details of the RF circuit are discussed in Section IV.

The +3 dBm IF signal from the main IF amplifier is amplified by a driver amplifier attached to the upconverter and applied to the diode through a low capacitance coaxial RF choke in the broad wall of the waveguide. The bandwidth of an upconverter is limited by the IF input circuit which typically has a 3 dB bandwidth of about 10 per cent. Analysis of the IF circuit including the effect of the frequency variable input resistance shows that if the varactor is driven from a constant resistance source the frequency variation in mismatch loss will compensate for the frequency variation in upconverter gain. Consequently the driver amplifier-upconverter interface is purposely mismatched to get the required 40 per cent bandwidth at the cost of reduced IF-to-RF gain. The IF circuit analysis and the amplifier performance is discussed in Section V.

Complete data on the upconverter performance is given in Section VI and summarized in Table I. The operating temperature range is −40°F to +140°F with small changes in the frequency response as shown in Fig. 2. Figure 3 is a photograph of the upconverter.

TABLE I—10.960 GHz BROADBAND UPCONVERTER PERFORMANCE

Pump power into filter	+20 dBm
IF power into amplifier	+3 dBm
Output power at band center	+16.2 dBm at +140°F and 76°F
	+16.9 dBm at −40°F
Frequency response at 76°F	1.4 dB down at 10.900 GHz
	1.0 dB down at 11.020 GHz
	ripples <0.2 dB peak-to-peak

Diode: Bell Telephone Laboratories L2280 with $V_B = 32$ V, $C_o = 1.091$ pF, $f_{co} = 137$ GHz.

III. MAXIMUM EFFICIENCY OPERATION AND DIODE SELECTION

If the output power is to be maximum with a given pump power, the upconverter must operate at maximum pump efficiency. In this section the operating point for maximum efficiency and the diode parameters are derived using the large signal analysis of Penfield and Rafuse, assuming a fully driven abrupt-junction silicon diode with $S_{\min} = 0$ and output load tuned.[4]

From Penfield and Rafuse, the pump efficiency, ϵ, and upper-sideband power, P_u in milliwatts, are

$$\epsilon = \frac{f_u}{f_p} \frac{m_s - \dfrac{f_u}{f_c}\dfrac{m_u}{m_p}}{m_s + \dfrac{f_p}{f_c}\dfrac{m_p}{m_u}}, \tag{1}$$

and

$$P_u = 16\pi(V_B + \Phi)^2 f_c C_{\min}\left(\frac{f_u}{f_c} m_s m_p m_u - \frac{f_u^2}{f_c^2} m_u^2\right), \tag{2}$$

where

f_s, f_p, f_u are the signal (IF), pump, and upper-sideband frequencies in GHz, and are known constants in this analysis,

m_s, m_p, m_u are the magnitudes of the normalized elastance coefficients,

f_c is the diode cutoff frequency in GHz defined as

$$f_c = \frac{10^3}{2\pi R_s C_{\min}} \quad \text{GHz}, \tag{3}$$

V_B is the diode reverse breakdown voltage,

Φ is the contact potential (0.7 volts for silicon),

and

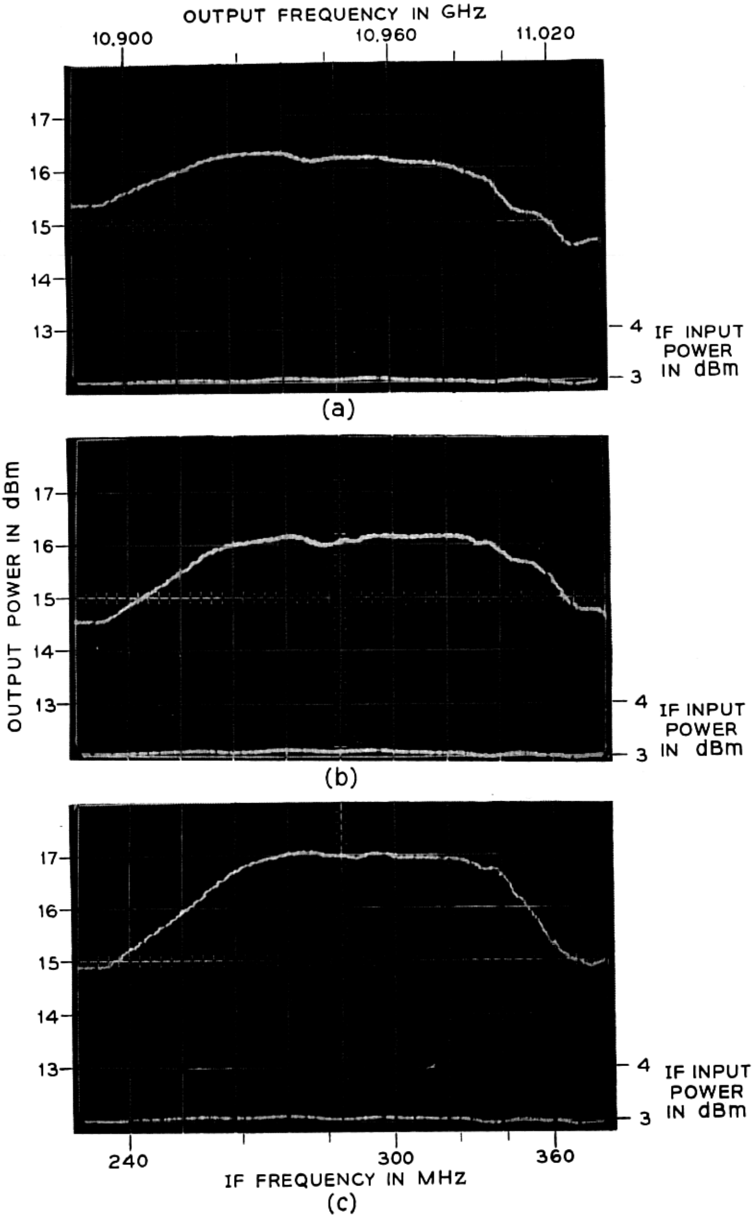

Fig. 2 — Upconverter frequency response at (a) +140°F, (b) +76°F, and (c) −40°F, with +3 dBm IF input power and +20 dBm pump power.

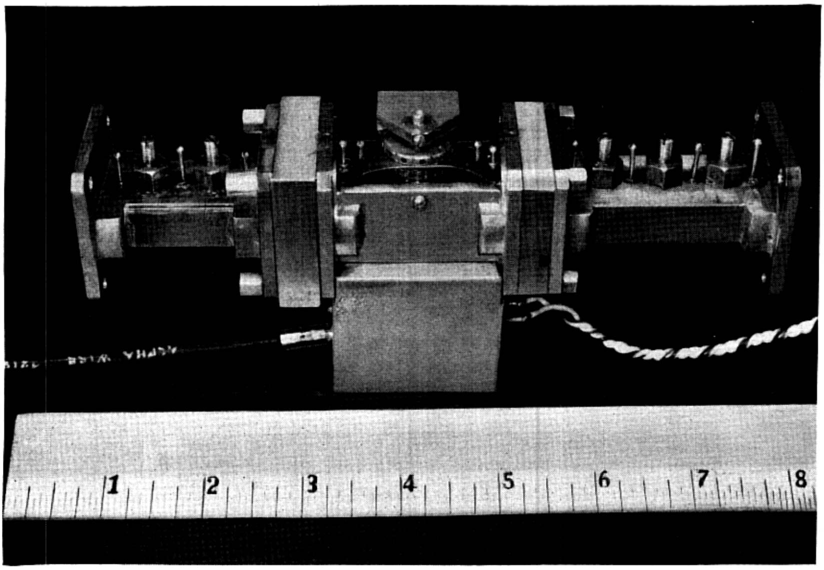

Fig. 3 — 10.960 GHz upconverter.

$C_{\min}$ is the minimum or breakdown capacitance in picofarads. For finite f_c and fully driven operation, that is,

$$m_s + m_p + m_u = 0.25, \tag{4}$$

ϵ has an absolute maximum when $m_s = 0.25$ and $m_u = m_p = 0$, a useless condition because the output power is zero. For $m_s < 0.25$, the ratio m_u/m_p which gives maximum efficiency can be found by differentiating (1). The result is

$$R = \left.\frac{m_u}{m_p}\right|_{\substack{\max \\ \mathrm{eff}}} = \frac{1}{m_s}\frac{f_p}{f_c}\left[\left(\frac{f_c^2}{f_p f_u}\, m_s^2 + 1\right)^{\frac{1}{2}} - 1\right]. \tag{5}$$

when f_c and m_s are given, m_p and m_u can be calculated from (4) and (5):

$$m_p = \frac{1}{R + 1}(0.25 - m_s) \tag{6}$$

$$m_u = \frac{R}{R + 1}(0.25 - m_s). \tag{7}$$

Substitution of (5), (6), and (7) into (2) eliminates m_u and m_p leaving only m_s, f_c, V_B, and $C_{\min}$. For any set of these variables the pump efficiency is maximum with respect to m_p and m_u.

In practice the diode parameters f_c, V_B, and $C_{\min}$ are not independent because of the experimentally determined dependence of breakdown voltage on material resistivity. Irvin has calculated f_c and f_{co}, the zero bias cutoff frequency, as a function of V_B for abrupt-junction silicon diodes using easily realizable breakdown versus resistivity values.[5] His curve for f_{co} agrees well with a curve by Lee giving the state of the art for epitaxial diffused silicon diodes (see Fig. 3).[6]

The equation,

$$f_c = \frac{14 \times 10^3}{C_o^{\frac{1}{6}}(V_B + \Phi)} \qquad \text{GHz}, \tag{8}$$

where C_o is in picofarads, closely approximates Irvin's calculated values of f_c versus V_B as shown in Fig. 4. The C_o dependence was obtained by plotting $1/f_c$ versus C_o for constant V_B for several commercially available diodes. Also shown in Fig. 4 are the f_{co} and V_B for several 1.0 pF Bell Telephone Laboratories L2280 silicon diodes obtained for this and related experiments.

Substituting (8) into (2), using the relation

$$C_{\min} = C_o \frac{\Phi^{\frac{1}{2}}}{(V_B + \Phi)^{\frac{1}{2}}}, \tag{9}$$

and setting $\Phi = 0.7$ gives an equation for P_u normalized to $C_o^{\frac{1}{2}}$:

$$\frac{P_u}{C_o^{\frac{1}{2}}} = \frac{6.97 \times 10^7}{f_c^{\frac{1}{2}}} \left(\frac{f_u}{f_c} m_s m_p m_u - \frac{f_u^2}{f_c^2} m_u^2 \right). \tag{10}$$

The following equations can be derived from the equations in Penfield and Rafuse in a similar way. In maximum pump efficiency operation, as defined by equation (5), the load and pump input resistances, R_u and $R_{\text{in},p}$, are equal.

$$\frac{P_{\text{in},p}}{C_o^{\frac{1}{2}}} = \frac{6.97 \times 10^7}{f_c^{\frac{1}{2}}} \left(\frac{f_p}{f_c} m_s m_p m_u + \frac{f_p^2}{f_c^2} m_p^2 \right). \tag{11}$$

$$\frac{P_{\text{in},s}}{C_o^{\frac{1}{2}}} = \frac{6.97 \times 10^7}{f_c^{\frac{1}{2}}} \left(\frac{f_s}{f_c} m_s m_p m_u + \frac{f_s^2}{f_c^2} m_s^2 \right). \tag{12}$$

$$C_o^{7/6} R_u = C_o^{7/6} R_{\text{in},p} = \frac{22.5 \times 10^3}{f_c^{\frac{3}{2}}} \left(\frac{f_c}{f_p} \frac{m_s m_u}{m_p} + 1 \right). \tag{13}$$

$$C_o^{7/6} R_{\text{in},s} = \frac{22.5 \times 10^3}{f_c^{\frac{3}{2}}} \left(\frac{f_c}{f_s} \frac{m_p m_u}{m_s} + 1 \right). \tag{14}$$

All these equations are valid for maximum pump efficiency operation of fully driven abrupt junction silicon diodes and any frequencies.

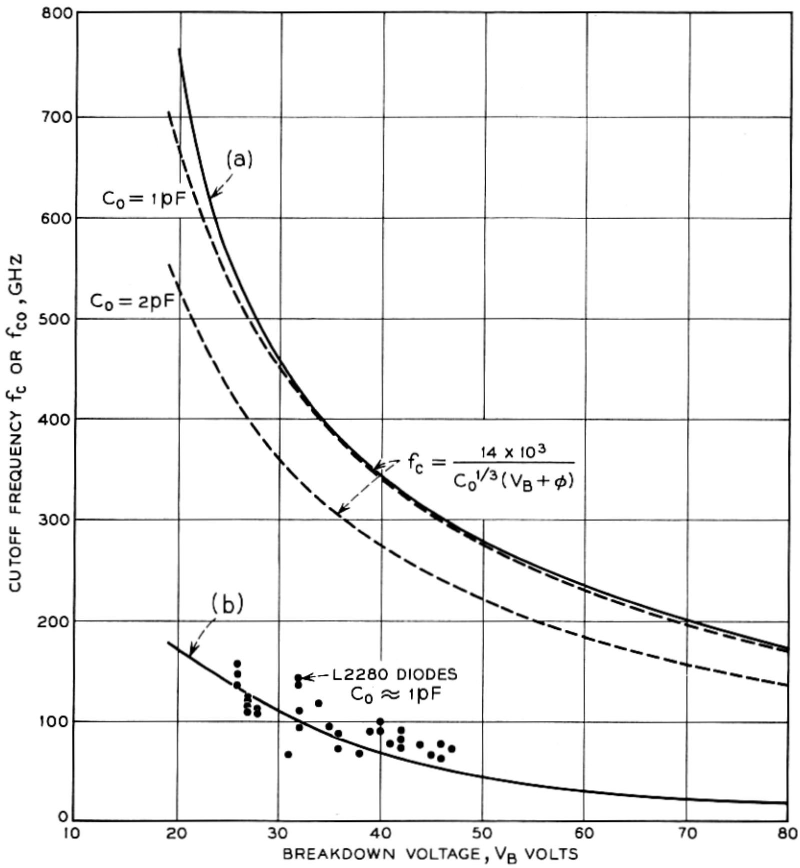

Fig. 4 — Cutoff Frequencies, f_c and f_{co}, versus breakdown voltage for silicon abrupt junction diodes: (a) calculated f_c for silicon abrupt junction (Ref. 5), $C_o = 1$ pF, (b) calculated f_{co} for silicon abrupt junction (Ref. 5), $C_o = 1$ pF and state of art for epitaxial diffused silicon (Ref. 6), $C_o = 0.8$ pF.

In Figs. 5, 6, and 7 the equations are plotted as a function of m_s with f_c as a parameter with $f_s = 300$ MHz, $f_p = 10.660$ GHz, and $f_u = 10.960$ GHz. The ratio R is plotted in Fig. 8 for use when calculating the values of m_u and m_p.

In these equations m_s is used as an independent variable even though it is not really independent but rather an intermediate variable which relates the independent variables, power and resistance. In Fig. 5 the powers are plotted as functions of m_s so that the relations between them are shown. Any point in Fig. 5 defines an operating point in the sense that if the indicated diode (defined by f_c), $P_{in,s}$, and $P_{in,p}$ are used and the diode terminated in the resistances shown in Fig. 7, the output power theoretically will be that value shown on the ordinate of Fig. 5 and m_s will have the value shown on the abscissa of Fig. 5. The values of m_s and f_c are used to locate the operating point on the graphs in Figs. 6, 7, and 8.

Even though the ratio of m_u to m_p was chosen to give maximum pump efficiency it is still possible to choose the remaining variables to get further maximization of pump efficiency or other optimizations. The required operating point can be found from Fig. 5 as follows.

The solid lines are lines of constant f_c and show the variation of normalized output power with m_s for a given diode. The peaks show that for a given diode maximum output power would be obtained by operating near $m_s = 0.10$.

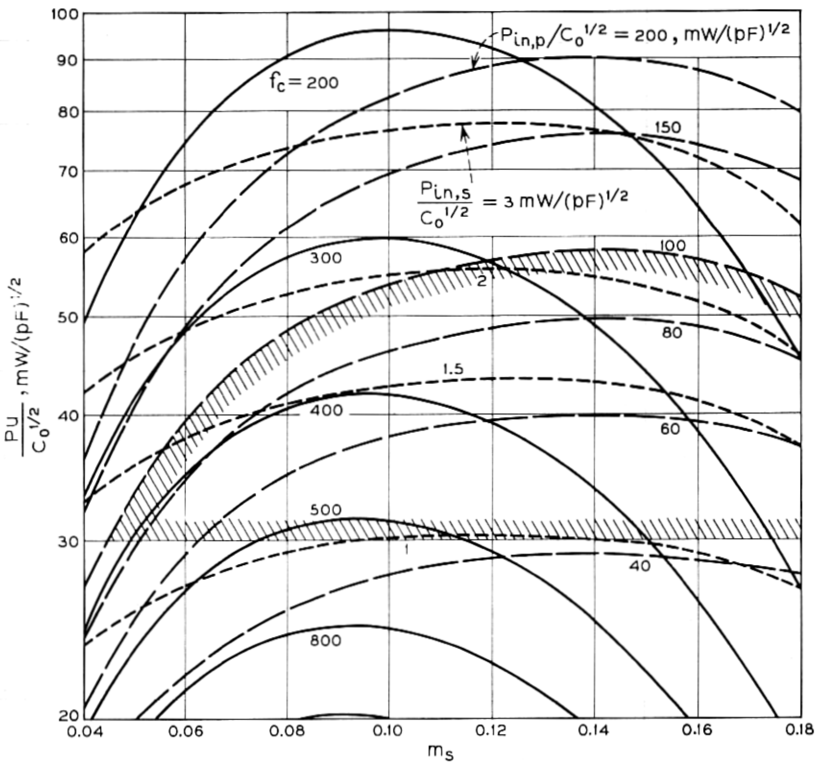

Fig. 5 — Upper-sideband output power versus m_s with f_c as a parameter and with contours of constant pump and signal power superimposed.

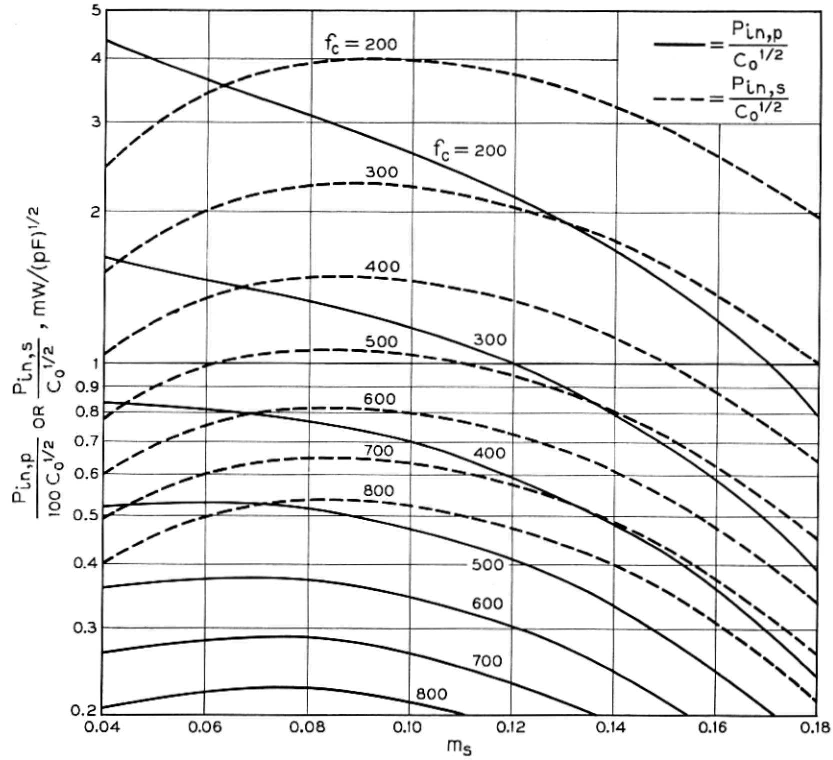

Fig. 6 — Normalized pump and signal power versus m_s with f_c as a parameter.

The broken lines are contours of constant normalized pump power. The peak in these curves shows that a given output power would be obtained with the least amount of pump power by operating near $m_s = 0.14$. Thus, for a given output power, kept constant by letting f_c vary, this is a point of maximum pump efficiency. However if f_c is constant the pump efficiency increases monotonically if the operating point is chosen at a larger m_s and the output power goes to zero.

The dashed lines are contours of constant normalized signal power. The peak in these curves shows that a given output power would be obtained with the least amount of signal power by operating near $m_s = 0.12$. Thus for a given output power this is a point of maximum gain. If f_c is kept constant, maximum gain occurs when operating near $m_s = 0.15$.

Figure 5 can also be used to define the set of operating points which meet the power requirements of a given problem. Using the radio system upconverter as an example, the shaded area defines a set of operating points each satisfying the requirements, $P_u \geqq 30$ mW and $P_{in,p} \leqq 100$ mW, when the diode capacitance is 1.0 pF. From Figs. 4 through 7 a theoretical operating point and diode were chosen. Since the output power theoretically obtainable is safely above the minimum needed, a lower value of m_s was chosen to avoid large values of R_u and $R_{in,p}$. The theoretical diode characteristics and upconverter diode performance are summarized in Table II. The average junction capacitance,

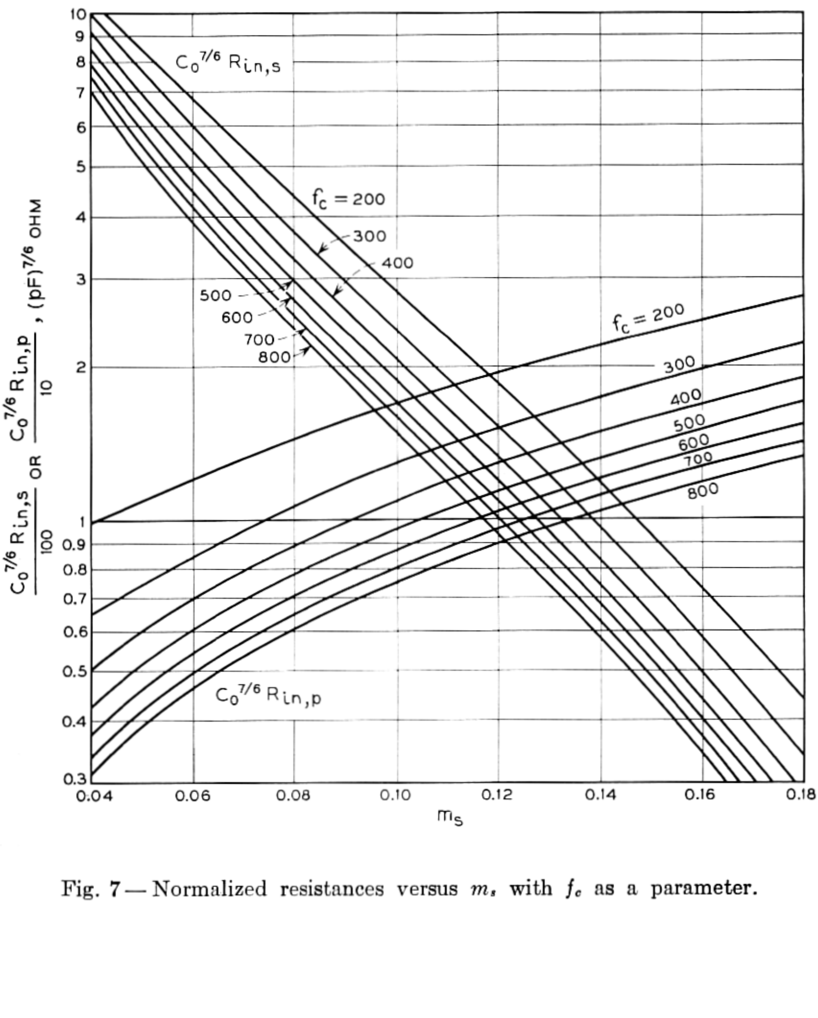

Fig. 7 — Normalized resistances versus m_s with f_c as a parameter.

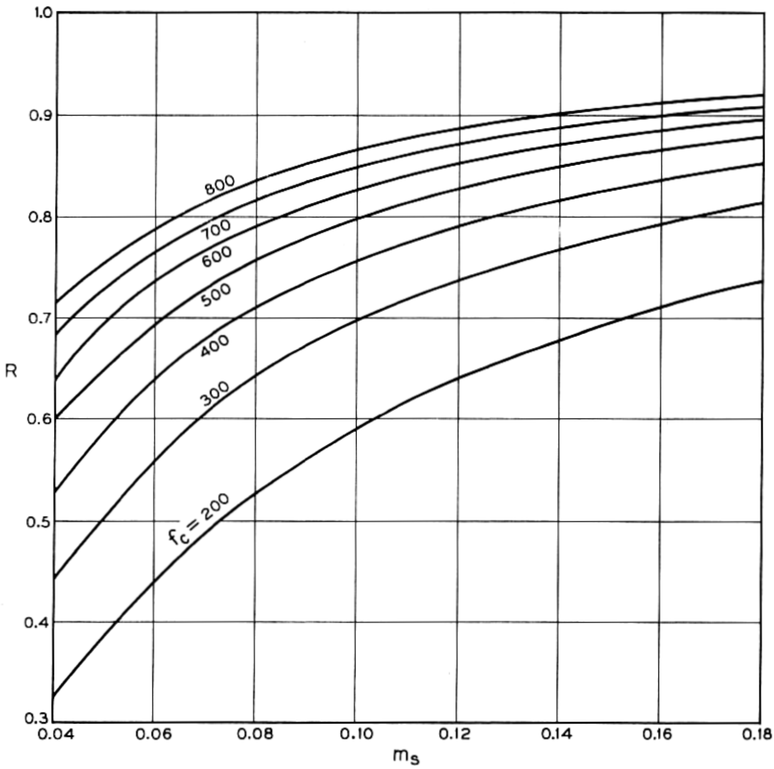

Fig. 8 — Plot of R, the ratio of m_u to m_p, versus m_s with f_c as a parameter.

C_j , is calculated from

$$C_j = 2C_{\min} = 2C_o\left(\frac{\Phi}{V_B + \Phi}\right)^{\frac{1}{3}} \text{pF} \tag{15}$$

assuming current pumping and cubic capacitance variation.*

IV. WAVEGUIDE CIRCUIT

Figures 9a and b show a cross section of the L2280 U-package diode and its equivalent circuit.[6] Above about 7 GHz the diode operates

* The actual capacitance variation of the L2280 diode is approximately cubic. Even though a square root variation was assumed for ease of analysis, use of a cubic variation in calculating C_j should give better agreement between theory and practice.

TABLE II—THEORETICAL UPCONVERTER DIODE CHARACTERISTICS AND PERFORMANCE

$f_c = 350$ GHz	$R_u = R_{in,p} = 10.6$ ohms	$P_{in,s} = 1.8$ mw
$V_B = 39$ volts	$R_{in,s} = 280$ ohms	$\epsilon = 52.4\%$
$f_{co} = 75$ GHz	$P_u = 49.2$ mw	$G = 27.2 = 14.3$ dB
$C_o = 1.0$ pF	$P_{in,p} = 93.9$ mw	$C_j = 0.521$ pF
$m_s = 0.09$		

above its self-resonant frequency and appears as a frequency dependent inductance in series with $R_{\text{in},\,p}$ as shown in Fig. 9c, where

$$L(\omega) = L_s\left(1 - \frac{1}{\omega^2 L_s C_j}\right). \tag{16}$$

At ω_o the equivalent parallel resistance and inductance, Fig. 9d, are

$$R_p(\omega_o) = R_{\text{in},p}\left[1 + \frac{\omega_o^2 L_s^2}{R_{\text{in},p}^2}\left(1 - \frac{1}{\omega_o^2 L_s C_j}\right)^2\right], \tag{17}$$

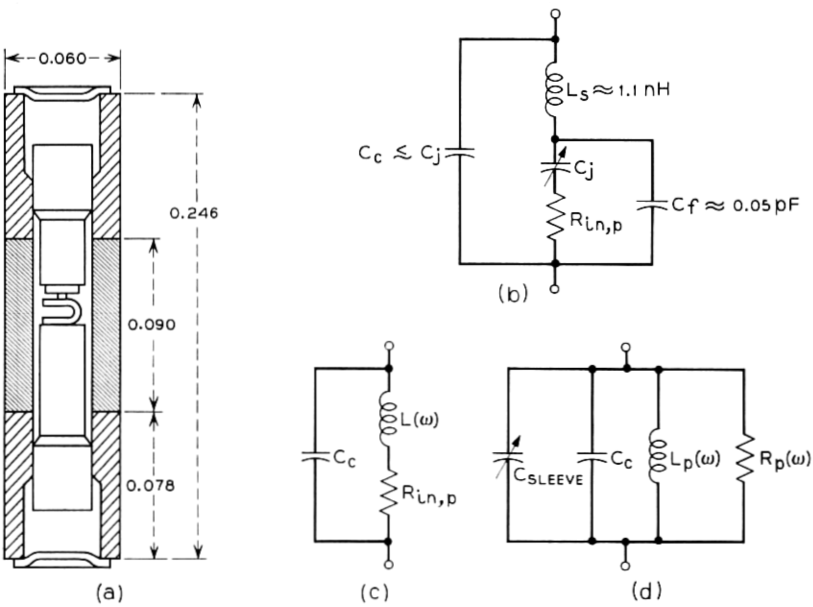

Fig. 9 — (a) L2280 U-package epitaxial diode (Ref. 6). (b) Equivalent circuit of diode. (c) Equivalent circuit of diode at 10.7 GHz. (d) Parallel equivalent of (c) with adjustable capacitance of the sleeve in parallel.

and

$$L_p(\omega_o) = L_s\left(1 - \frac{1}{\omega_o^2 L_s C_j}\right)\left[1 + \frac{1}{\dfrac{\omega_o^2 L_s^2}{R_{\text{in},p}^2}\left(1 - \dfrac{1}{\omega_o^2 L_s C_j}\right)^2}\right]. \tag{18}$$

The total capacitance in parallel with the diode, including C_c, required to tune $L_p(\omega_o)$ at ω_o is

$$C_T = \frac{L_s\left(1 - \dfrac{1}{\omega_o^2 L_s C_j}\right)}{R_{\text{in},p}^2 + \omega_o^2 L_s^2\left(1 - \dfrac{1}{\omega_o^2 L_s C_j}\right)^2}. \tag{19}$$

Substituting the values of $R_{\text{in},\,p}$ and C_j from Table II and $L_s = 1.1$ nH and $f_o = 10.900$ GHz into equations (17) and (19), gives $R_p =$ 222 ohms and $C_T = 0.294$ pF. This value of C_T can be easily obtained by adding a small capacitance, C_{sleeve}, in parallel with the diode, thus using the diode self-inductance and parallel capacitance as a tuned impedance transformer. The R_p of 222 ohms across a waveguide with $Z_o = 420$ ohms can be matched easily with three $\lambda/8$ tuning screws. This method of impedance matching eliminates the need for stepped impedance transformers and results in a significant simplification of the RF circuit.

Physically, the diode is mounted across a gap formed by two cylindrical posts protruding into standard height X-band waveguide from opposite broad walls; see Fig. 10. The lower post is a diode chuck and also the center conductor of the coaxial RF choke through which the IF signal is applied. The top post is a diode chuck surrounded by an adjustable cylindrical sleeve. The diode tuning capacitor is formed by the end of the sleeve and the lower diode chuck. The capacitance can be varied from approximately 0.1 pF to some large value when the sleeve touches the lower chuck.

The Q and the bandwidth of the RF impedance matching circuit can be calculated from (17) and (19):

$$Q(f_o) = 2\pi f_o C_T(f_o) R_p(f_o) = 4.47.$$

Ignoring the frequency dependence of $R_{\text{in},\,p}$ and R_p, the 3 dB bandwidth is approximately

$$BW = \frac{f_o}{Q(f_o)} = 2.4 \text{ GHz},$$

which is much larger than the required 120 MHz. Therefore the diode will be matched to the waveguide over the 120 MHz band and the RF bandwidth will be determined by the output filter.

The pump and upper-sideband output filters are direct coupled filters with multiple posts as the shunt inductive reactances. A two-section maximally flat filter is used for the pump filter and a three-section 0.1 dB Tchebyscheff filter is used for the upper-sideband output filter. The insertion loss and return loss of the output filter are shown in Fig. 11; the measured performance of both filters is summarized in Table III.

V. IF CIRCUIT

The main consideration in the IF circuit is the bandwidth requirement. Since the required 120 MHz bandwidth is a very small percentage bandwidth at RF, the RF circuit presents no problem as far as bandwidth is concerned. At the 300 MHz IF, however, the 120 MHz represents a 40 percent bandwidth. An estimate of the severity

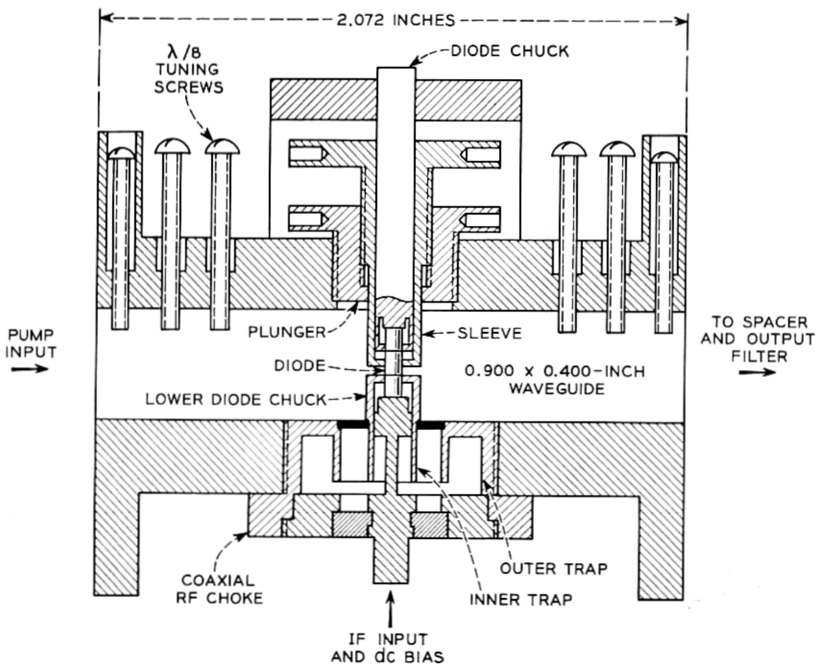

Fig. 10 — Cross section of upconverter.

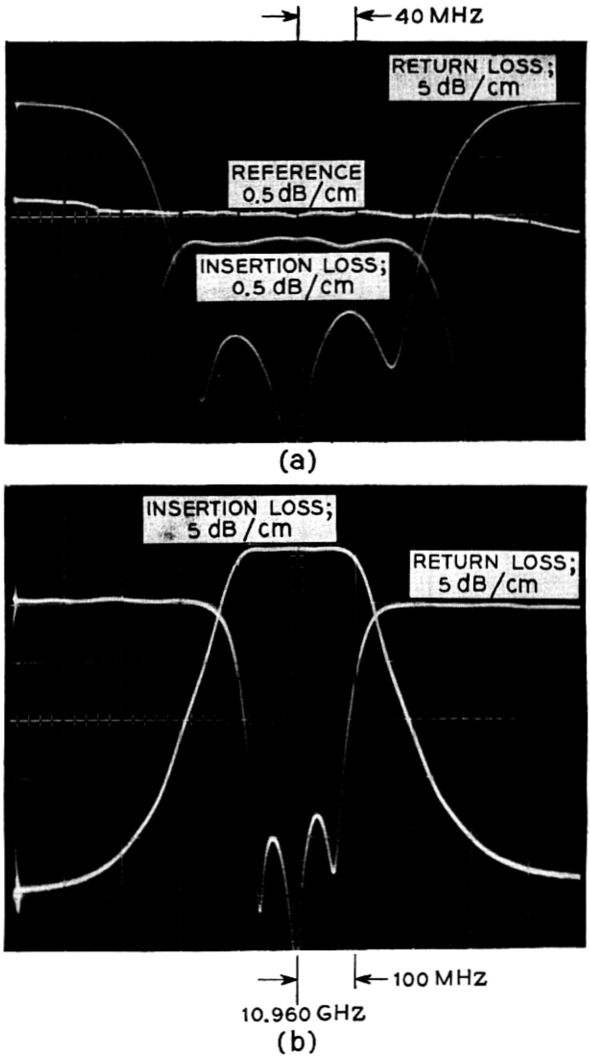

Fig. 11 — Return loss and insertion loss of the 3 section 0.1 dB Tchebyscheff output filter.

TABLE III—MEASURED PERFORMANCE OF PUMP AND OUTPUT FILTERS

	Pump filter	Upper sideband output filter
Type	Maximally flat	0·1 dB Tchebyscheff
Number of sections	2	3
Center frequency	10.658 GHz	10.960 GHz
Bandwidth	134 MHz at 3 dB	152 MHz at 0.1 dB
Skirt attenuation	20 dB at 10.900 GHz	26.5 dB at 10.660
Midband loss	0.3 dB	0.2 dB
Return loss	27 dB at 10.658 GHz	>18 dB in pass band

of the problem can be found by calculating the Q of the IF input circuit shown in Fig. 12. The input impedance at the RF choke terminals is

$$Z = \frac{R_{\text{in},s}}{\left(1+\frac{C_s}{C_j}\right)^2\left[1+\omega^2 R_{\text{in},s}\left(\frac{C_sC_j}{C_s+C_j}\right)^2\right]} + \frac{\left[1+\omega^2 R_{\text{in},s} C_j\left(\frac{C_sC_j}{C_s+C_j}\right)\right]}{j\omega(C_s+C_j)\left[1+\omega^2 R_{\text{in},s}\left(\frac{C_sC_j}{C_s+C_j}\right)^2\right]} \tag{20}$$

where the frequency variation of $R_{in,\,s}$ has been neglected. For the values given in Table II, the impedance is approximately

$$Z = \frac{R_{\text{in},s}}{\left(1+\frac{C_s}{C_j}\right)^2} + \frac{1}{j\omega(C_s+C_j)}. \tag{21}$$

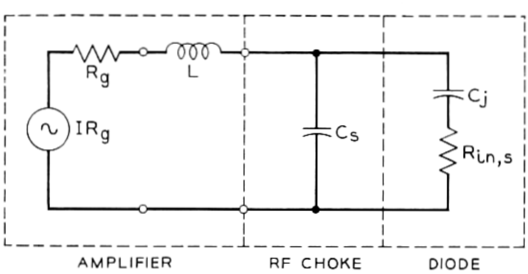

Fig. 12 — IF input circuit.

The Q and 3 dB bandwidth at $f_o = 300$ MHz are

$$Q = \frac{\left(1 + \dfrac{C_s}{C_j}\right)^2}{\omega_o R_{\text{in},s}(C_s + C_j)} = 15.2$$

$$BW = \frac{2\pi f_o^2 R_{\text{in},s}(C_s + C_j)}{\left(1 + \dfrac{C_s}{C_j}\right)^2} = 19.8 \text{ MHz}.$$

Simple resistance loading to lower the Q would require a total resistance about 6 times the unloaded input resistance, would result in 15.7 dB of loss, and the frequency response would still be down 3 dB at the band edge. Some improvement on this loss could be achieved by using a more elaborate filter, but the filter itself could contribute considerable loss.

The inherent narrow bandwidth is caused by the shunt capacitance, C_s, which limits the high frequency response and the junction capacitance, C_j, which in turn limits the low frequency response giving a bandpass type of characteristic. To widen the bandwidth, C_s should be smaller and C_j larger. The value of C_s is primarily determined by the capacitance of the RF choke which is required to isolate the IF circuit from the RF. Considerable effort was required to design a choke with a capacitance of 1.23 pF; it is not likely that this capacitance can be reduced significantly. To increase C_j would require a higher capacitance diode resulting in a complete redesign of the RF circuit. Therefore the situation cannot be improved easily by changing the circuit parameters.

The solution is in the inverse frequency variation of $R_{\text{in},s}$, equation (14), and upconverter IF gain, the ratio of equation (10) to (12). Neglecting L, the power transfer of the input circuit of Fig. 12 with $R_{\text{in},s}$ constant is a bandpass characteristic with band-edge slopes of 6 dB per octave. The frequency variation of $R_{\text{in},s}$ causes the slopes to be only 3 dB per octave. If the IF band lies on the low frequency side of the bandpass characteristic, the 3 dB per octave increase in power transfer will just compensate for the 3 dB per octave decrease in IF gain giving an overall constant gain relative to the available power from the generator. Since the IF band must lie below the peak of the power transfer characteristic there must be some mismatch loss. However, the bandwidth can be very wide and is limited practically by the available power from the generator and filter requirements.

The equation for the power into $R_{\text{in},s}$ relative to the generator available power is derived in the appendix and the response in dB is plotted in Fig. 13 as a function of normalized frequency, $\omega/r\omega_o$, for several representative combinations of C_j and L. The diode and operating point is that given in Table II. The location of a 120 MHz band centered at 300 MHz with $r = 5$ is shown by the shaded area. The inductance L is used to decrease the loss at the peak of the response and to present an open circuit to frequencies in the diode which are below the rejection band of the RF choke. The value of L cannot be too large or it will resonate with the equivalent input capacitance and cause a very narrow band response as shown in Fig. 13 for $L = 80$ nH. Figure 13 also shows that the amount of mismatch loss is determined by the value of C_j . The location of the power transfer peak is determined by the value of the inductance, L, and the shunt capacitance, C_s , the latter being determined primarily by the choke capacitance.

The IF signal incident at the choke terminal of Fig. 12 must be large enough to supply the required power to the diode through the mismatched input circuit and must be supplied by a source of about 50 ohms impedance. The source must be physically close to the choke to minimize stray reactances because of the relatively high input frequency. The return loss requirement on the transmitter input

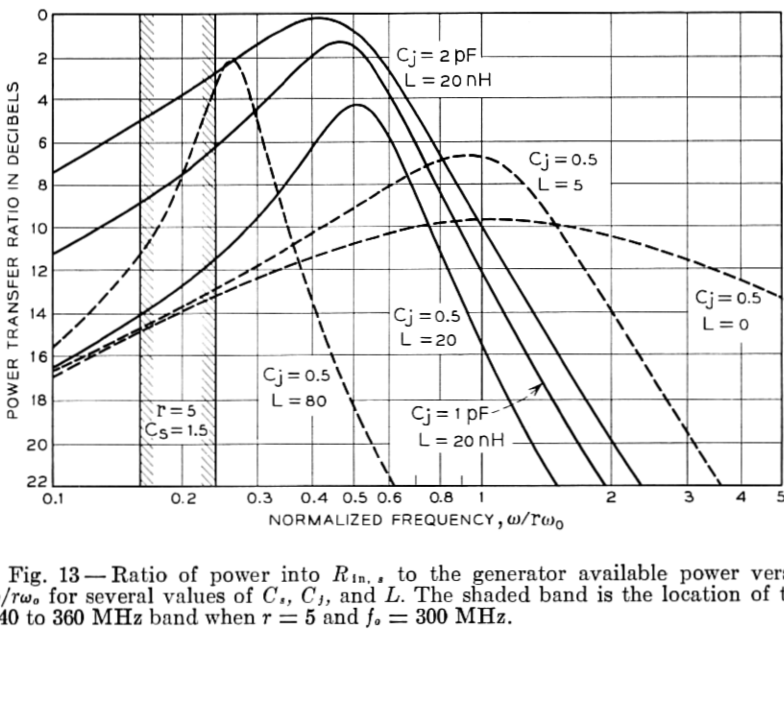

Fig. 13 — Ratio of power into $R_{in,s}$ to the generator available power versus $\omega/r\omega_o$ for several values of C_s, C_j, and L. The shaded band is the location of the 240 to 360 MHz band when $r = 5$ and $f_o = 300$ MHz.

in the short hop radio system experiment is approximately 18 dB over the IF band to prevent echoes in the long cable between the receiver and transmitter. All these requirements are met by using an IF driver amplifier as an integral part of the upconverter.

Figure 14 is a schematic diagram of the driver amplifier. Two Western Electric 45B transistors, used as common-base transformer coupled stages, are required because of the relatively high output power and frequency. With 0 dB input power the amplifier has 6.5 dB gain between 50 ohm terminations; the gain-frequency response is down 0.4 dB at 230 MHz and 360 MHz as shown in Fig. 15a. The input return loss, shown in Fig. 15b, is larger than 20 dB from 200 to 400 MHz. The output return loss, Fig. 15c, is 17 dB at 340 MHz and drops to 3 dB at 240 MHz as expected from the single-tuned output circuit.

VI. UPCONVERTER MEASUREMENTS

Initial measurements were made on the 10.760 GHz upconverter using a constant 300 MHz IF signal of +5 dBm from a 50 ohm source. The IF signal was applied to a coupling capacitor, inductor, and bias circuit which were similar to C_8, L_3, and the bias circuit of Fig. 14. Then the RF circuit was tuned for maximum output power. The actual diode performance, given in Table IV compares well with the the-

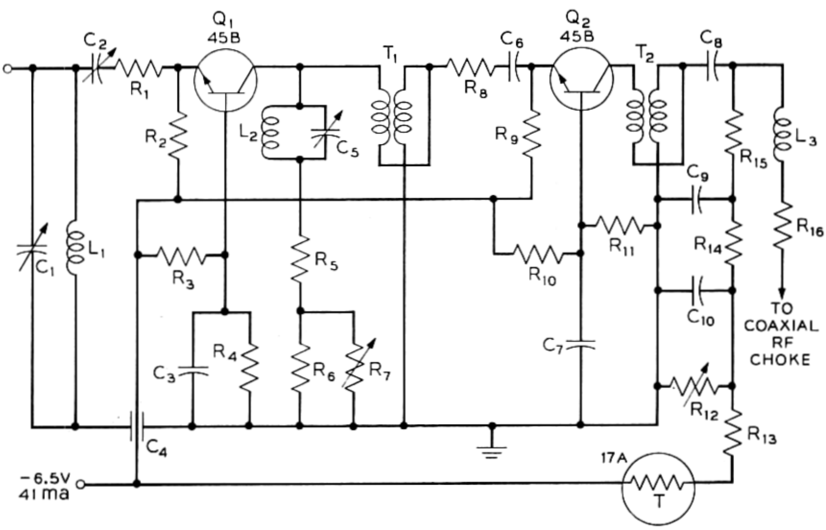

Fig. 14 — Schematic diagram of upconverter driver amplifier.

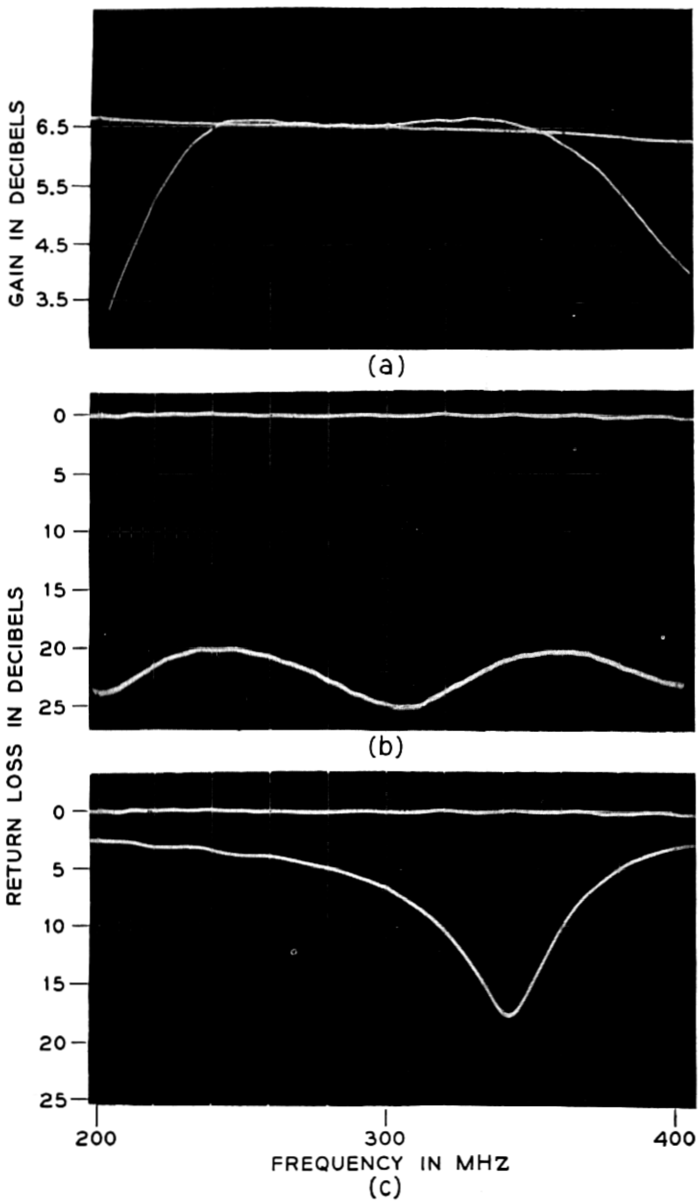

Fig. 15 — IF driver amplifier (a) insertion gain, (b) input return loss, and (c) output return loss versus frequency.

TABLE IV—10.760 GHz UPCONVERTER PERFORMANCE AND DIODE CHARACTERISTICS WITH A CONSTANT 300 MHz IF SIGNAL

$V_B = 32$ V	$f_u = 10.760$ GHz	$P_{in,s} = 2.5$ mw (+4 dBm)
$f_{co} = 112$ GHz	$f_s = 300$ MHz	$\epsilon = 44\%$
$C_o = 1.075$ pF	$P_u = 42$ mw	$G = 12.3$ dB
$f_p = 10.460$ GHz	$P_{in,p} = 96$ mw	

oretical performance in Table II. The output power was maximum when the diode was self-biased at about 1.0 volt. The low bias voltage indicates that the diode is being overdriven causing its average junction capacitance to be larger than the theoretical value which, from Fig. 13, accounts for the low IF circuit mismatch loss. This initial performance showed that the 9.5 dBm available power from the driver amplifier should be sufficient even for broadband operation.

Since the upconverters were intended to work directly into an antenna, the effect of a poor antenna return loss on the upconverter was measured on the 10.760 GHz upconverter by terminating it in a variable attenuator followed by a sliding short. To determine the effect of the phase of the reflection, several sweeps were recorded on the same photograph as the position of the sliding short was changed. The result is an envelope of the possible gain-frequency responses for each return loss magnitude. The envelopes for a return loss of > 23, 20, 10, and 3 dB are shown in Fig. 16. The upconverter was stable and did not produce extraneous tones for any magnitude or phase of reflection.

Figures 17a and b show the variation of output power and frequency response as the pump and IF levels were varied with the upconverter at 76°F. In Fig. 17a the IF power incident on the driver amplifier was kept at +3 dBm and the pump power was varied in 1 dBm steps from +16 dBm to +22 dBm without retuning. There is very little compression and little change in band shape. In Fig. 17b the pump power was kept at +20 dBm and the incident IF power was varied in 1 dB steps from −3 dBm to +4 dBm without retuning. There is considerable compression and change in band shape. This shows that the diode is being driven very hard by the IF signal and in a sense is being pumped by the IF signal.

Figure 17c shows the return loss at both pump port and driver amplifier input port as the input frequency was swept. The pump return loss does not have the conventional meaning because the pump signal was not swept; it was constant in frequency, the pump

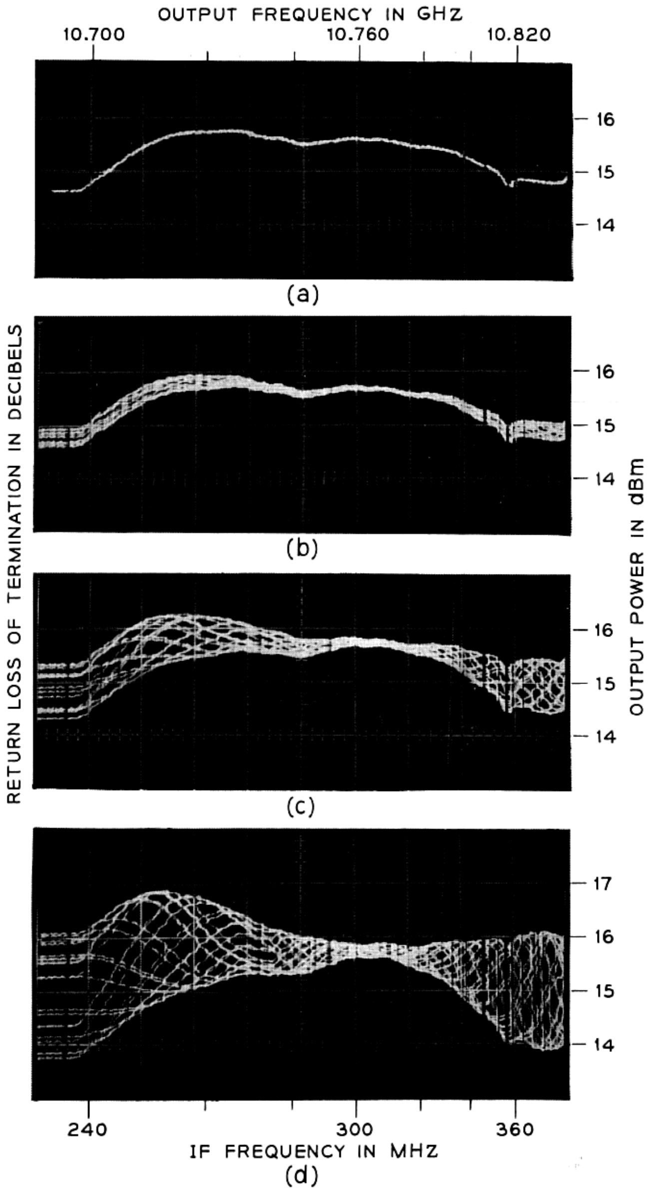

Fig. 16 — Frequency response of the upconverter as the phase of the terminating return loss is varied for return loss magnitudes of (a) > 23 dB, (b) 20 dB, (c) 10 dB, (d) 3 dB.

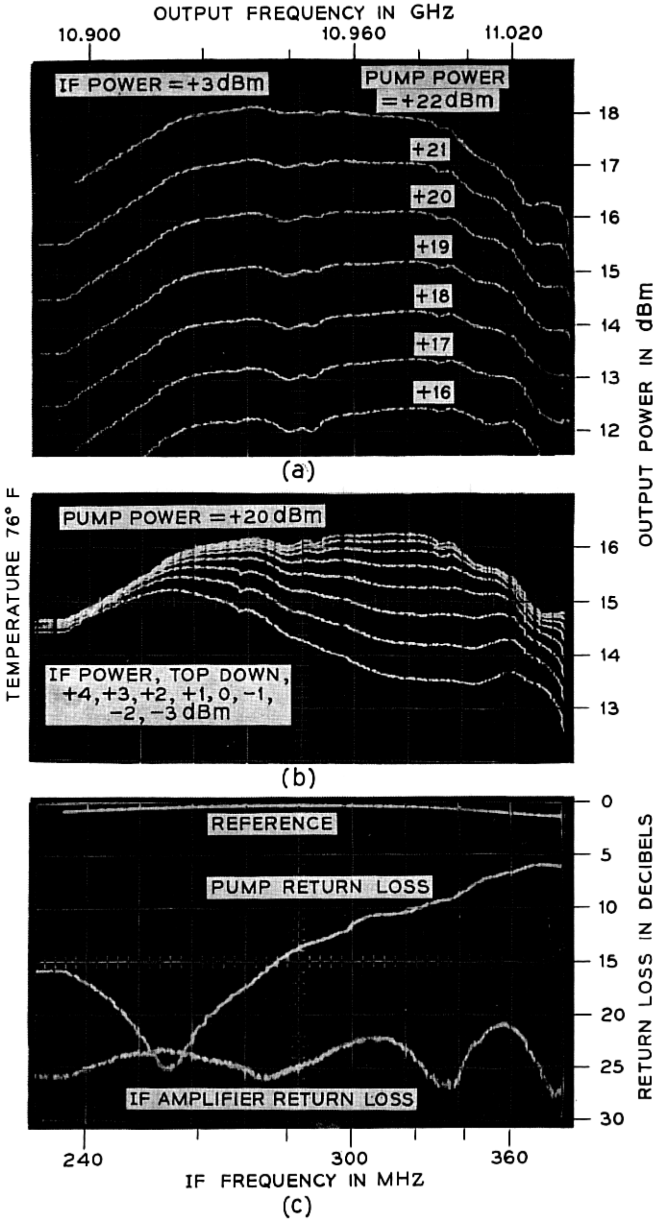

Fig. 17 — (a) Frequency response for different values of pump power. (b) Frequency response for different values of IF power. (c) Return loss at pump and IF amplifier input ports.

reflection varying because of the swept IF signal. The amplifier input return loss is not the same as in Fig. 15b because of the imperfect isolation between amplifier output and input, and because the output was terminated in the variable upconverter impedance rather than 50 ohms.

The diode bias was measured at the top of R14 in Fig. 14 with +20 dBm pump power and a +3 dBm IF signal at 300 MHz. The bias voltage was -1.5 volts and did not vary as the pump power varied but decreased 0.1 volt per dB as the incident IF power was decreased again indicating that the diode is being pumped primarily by the IF.

VII. DISCUSSION

Some bandshape and stability problems were caused by parametric oscillations in the IF circuit at frequencies below the RF choke rejection band and above the self-resonance of the series inductance in the IF amplifier output circuit. A pumped varactor diode can produce signals at any pair of frequencies whose sum is equal to the pumping frequency, if the terminating impedances are favorable at both frequencies. Since the RF choke rejects frequencies above about 10 GHz, reactive termination of all frequencies below 5 GHz reactively terminates one of the frequencies in any pair and prevents any parametric oscillation. However, this requires an inductance with a self-resonance above 5 GHz which is difficult to obtain. In retrospect a better solution would be to use a miniature 50 ohm low-pass filter, which has no spurious responses to above about 25 GHz, to replace both the inductance and RF choke. If the amplifier output were 50 ohms the filter could be considered part of the 50 ohm source and the IF circuit could be operated as discussed in Section V except that the major part of the shunt capacitance would be eliminated. The output element of the filter would have to be a shunt capacitor to provide an RF short circuit at the waveguide wall.

APPENDIX

Derivation of Power Transfer Characteristic of IF Input Circuit

Referring to Fig. 12, the available power from the generator is

$$P_A = \frac{I^2 R_g}{4}. \tag{22}$$

The power into $R_{\text{in},\,s}$ is

$$P_L = |\, I_2 \,|^2 \, R_{\text{in},s} \,, \tag{23}$$

where I_2 is the loop current flowing through $R_{\text{in},s}$ and is given by

$$I_2 = \frac{IR_g}{(1 - \omega^2 LC_s + j\omega C_s R_g)\left(R_{\text{in},s} + \dfrac{1}{j\omega C_s} + \dfrac{1}{j\omega C_j}\right) - \left(\dfrac{1}{j\omega C_s}\right)}. \tag{24}$$

The ratio of load power to available power is

$$\frac{P_L}{P_A} = \frac{4R_g R_{\text{in},s}}{\left|(1 - \omega^2 LC_s + j\omega C_s R_g)\left(R_{\text{in},s} + \dfrac{1}{j\omega C_s} + \dfrac{1}{j\omega C_j}\right) - \left(\dfrac{1}{j\omega C_s}\right)\right|^2}. \tag{25}$$

The frequency variation of $R_{\text{in},s}$ is given by equation (14). Since $f_c \gg f_s$,

$$R_{\text{in},s} \approx \frac{22.5 \times 10^3}{C_o^{7/6} f_c^{1/2}} \frac{m_p m_s}{m_u} \left(\frac{1}{f}\right)\left(\frac{f_o}{f_o}\right) = \frac{R_{\text{in},s}(\omega_o)}{\omega/\omega_o}, \tag{26}$$

where ω is the variable IF and ω_o is the center frequency of the IF band. From Table II the value of $R_{\text{in},s}(\omega_o)$ is 280 ohms. Let the ratio of

$$R_{\text{in},s}(\omega) = \frac{\omega_o}{\omega} rR_g . \tag{27}$$

Substituting (27) into (25) and taking the absolute value of the denominator gives

$$\frac{P_L}{P_A} = 4\frac{\omega}{r\omega_o} \Bigg/ \left\{\left[1 + \left(\frac{\omega}{r\omega_o}\right)\left(1 + \frac{C_s}{C_j}\right) - \left(\frac{\omega}{r\omega_o}\right)(r^2\omega_o^2 LC_s)\right]^2 + \left[\frac{1}{\omega_o r R_g C_j} - \frac{\omega}{r\omega_o}(r\omega_o R_g C_s) - \left(\frac{\omega}{r\omega_o}\right)^2 \frac{r\omega_o L}{R_g}\left(1 + \frac{C_s}{C_j}\right)^2\right]\right\}. \tag{28}$$

The following frequencies are defined and substituted into (28) to get (29):

$$\omega_L = \frac{R_g}{L}, \qquad \omega_c = \frac{1}{R_g C_s}, \qquad \omega_j = \frac{1}{R_g C_j}.$$

$$\frac{P_L}{P_A} = 4\frac{\omega}{r\omega_o} \Bigg/ \left\{\left[1 + \frac{\omega}{r\omega_o}\left(1 + \frac{C_s}{C_j}\right) - \left(\frac{\omega}{r\omega_o}\right)^2 \frac{r^2\omega_o^2}{\omega_L \omega_c}\right]^2 + \left[\frac{\omega_j}{r\omega_o} - \left(\frac{\omega}{r\omega_o}\right)\left(\frac{r\omega_o}{\omega_c}\right) - \left(\frac{\omega}{r\omega_o}\right)^2\left(\frac{r\omega_o}{\omega_L}\right)\left(1 + \frac{C_s}{C_j}\right)\right]^2. \tag{29}$$

Now let x be the frequency normalized to $r\omega_o$, for example,

$$x_L = \frac{\omega_L}{r\omega_o} = \frac{R_g}{r\omega_o L}$$

giving finally

$$\frac{P_L}{P_A} = \frac{4x}{\left[1 + x\left(1 + \frac{x_i}{x_c}\right) - x^2 \frac{1}{x_c x_L}\right]^2 + \left[x_i - \frac{x}{x_c} - \frac{x^2}{x_L}\left(1 + \frac{x_i}{x_c}\right)\right]^2}. \tag{30}$$

Equation (30) is plotted in dB in Fig. 13 as a function of the normalized frequency x for several representative combinations of C_j and L; r is assumed to be 5 which approximates the ratio of $R_{in,s}(\omega_o) = 280$ ohms to a generator impedance of about 50 ohms.

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