THE TRAVELING WAVE AMPLIFIER AS A BISTABLE OSCILLATOR

Linual Dale Smithey

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by

Linual Dale Smithey

Lieutenant Commander,

United States Naval Reserve

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ABSTRACT

Multi-mode oscillators have frequency memory, allowing information to be stored indefinitely. They permit destruct or non-destruct read-out. Systems having upward of a thousand modes are possible. Bistable oscillators using Traveling Wave Amplifiers are capable of counting rates greater than one hundred megacycles per second.

General requirements imposed upon a multi-mode oscillatory system are discussed, including mode determination and stability. Problems germane to adapting a short regenerative loop to a Traveling Wave Amplifier to provide bistable operation and very fast switching are considered. Application of multiple loop feedback principles is recommended for mode selection and improved mode stability. The short regenerative loop, including only the minimum tube length necessary for adequate gain, minimizes delay and speeds the transition.

Spurious feedback and difficulty in measuring operating conditions are the principle deterents in an S-band experiment. Inadvertent feedback was not eliminated, but was ultimately used to secure bistable oscillation. Stability was good. Conditions were not too critical.

Methods of exploiting beam modulation for mode switching are discussed. Two methods were tried, with inadequate equipment. Erratic switching was obtained with one (a heterodyne method.)

Traveling Wave Amplifiers show much promise in this application. Many problems remain to be solved. Principle among them is the control of regeneration. To this end, a tube with a helix only as

long as necessary to provide the requisite gain is recommended. Further, the unterminated ends of the helix can be used as reflection points. A single coupler will afford readout and three feedback loops with adjustable parameters. For an S-band tube, the helix only needs be about one and one-half inches long, entailing a loop delay of about six millimicroseconds. This should allow a counting rate near 150 mcs.

PREFACE

The tremendous advances in the fields of nuclear studies and high speed digital computers have created pressing demands for counting of random phenomena separated in time by a few milli-microseconds. Speeds of binary counters using conventional tubes and circuits are limited to about one hundred megacycles per second. It is reasonably evident that higher speeds will be attained using distributed amplifiers of large bandwidth, of which presently the Traveling Wave Amplifier is the principle exponent. This paper will survey briefly some principles admitting the use of the Traveling Wave Amplifier as a bistable oscillator, then describe an experimental study of some of its features that are of importance in adapting it to this purpose.

The writer wishes to express his gratitude to Messrs. Hewlett and Packard, in whose laboratories the experimental work was conducted. He is especially indebted to Dr. Peter D. Lacy, Group Leader for Microwave Tube and Equipment Development for the Hewlett-Packard Company, under whose supervision the work was performed, and with whom many of the ideas examined are original. Thanks are also due Dr. Wm. M. Bauer, Professor M. Pastel and Professor M. L. Cotton, of the U.S. Naval Postgraduate School, whose interest and professional counsel have been most valuable.

L. D. Smithey,

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TABLE OF SYMBOLS AND ABBREVIATIONS

(Listed in order of their use)

gm	grid-plate transconductance
C	capacity
RC	resistance-capacity product
t_r	rise time
mmf	10^{-12} farad
ma	milliampere
mμsec	millimicrosecond
mcs	megacycle per second
kcs	kilocycle per second
f	frequency
Q	a quality factory ($Q = f/\delta f$)
e	the base of natural logarithms
p	the complex variable
τ	a delay factor or rise time
μ	a gain factor
β	a feedback factor, bandwidth
e(t)	a function of time
E(p)	a function of complex frequency
$\ln$	natural logarithm
σ	the real portion of p
$j\omega$	the complex portion of p

α	an attenuation factor
ϕ	a space-phase angle
ω	$2\pi f$
δf	a frequency increment
Re	real part of
Im	imaginary part of
op. cit.	Opere Citato - in the work cited
TWO	Traveling Wave Oscillator
rf	radio frequency
$J_n(\Delta\phi)$	Bessel function of the first kind and order n with argument $\Delta\phi$
$\Delta\phi$	phase deviation
TWT	Traveling Wave Tube
TWA	Traveling Wave Amplifier
VSWR	Voltage Standing Wave Ratio
CW	Continuous Wave
db	decibel
dbm	decibel referred to one milliwatt in 50 ohms
kmcs	kilomegacycle per second
μsec	microsecond
coax	coaxial - coaxial line
V_h	helix voltage
I_b	beam current
k	coupler
ac	Alternating Current
dc	Direct Current

CHAPTER I

HIGH SPEED COUNTING

1. Introduction.

The tremendous advances in the fields of nuclear studies and the advent of high speed digital computers have created a pressing demand for counting of random phenomena separated in time by a few millimicroseconds. Active development of counting techniques has continued apace, but many demands remain unsatisfied by speeds presently attainable. We shall survey briefly the elements of binary counting and show that a fundamentally new approach is necessary if an appreciable increase in counting rate is to be achieved.

Features of counting systems that determine their suitability for application are memory, accuracy and speed of response. High speed memory systems are not required for counting periodic events, because frequency translation permits counting at essentially any desired rate. The memory feature is, however, essential to the counting of random phenomena, since a rate within the purview of normal definition is meaningless. This imposes the requirement of stability in state on the counter.

We are here concerned only with electronic counting devices, and the counting of phenomena for which electrical manifestations are available. Attention will be confined to the binary counting of electrical impulses suitable shaped, with emphasis on speed. No generality will be lost in considering only binary counters, since higher order systems can perform no faster than these.

2. Binary Counters - General Considerations.

The three characteristics of electronic systems that are most readily used for counting are energy level, voltage level and frequency. Methods of using energy and voltage levels are well developed, but the use of frequency memory is quite new, and does not yet compete in speed.

Placing increments of charge on a capacitor in response to each input pulse has been the most common method of counting by energy storage. This method is suitable for relatively slow rates. It has reliable memory for a comparatively short time due to capacitor leakage. It is, in general, not suited to binary counting. The reliability of read-out becomes uncertain above ten or twelve counts because of exponential charging unless special precautions are taken.

Counting by voltage states may be done in a variety of ways, but by far the most common one uses a direct-coupled bistable multivibrator. This circuit, often called the Eccles-Jordan circuit, appears in a variety of configurations employing triodes, pentodes or a combination of the two. The Phantastron family of circuits is used in some applications, but in high speed applications the multivibrator is invariably used. Woodbury and Holdam [3][1] give a good discussion of general purpose counters as of 1949, and some of the following considerations are drawn therefrom.

3. Speed Limitations in Lumped Constant Circuits.

Figure 1 shows the capacities that limit the speed of response of a typical bistable direct-coupled multivibrator. The effect of

1. Bracketed numbers refer to bibliography items.

grid-cathode capacity may be compensated in part, by use of a "speed-up" capacitor, but ultimately this capacity must be charged. The influence of Miller feedback through the plate-grid capacity is quite severe, too, and multi-grid tubes must be used for the highest counting rates. The principle improvement lies in the virtual elimination of the plate-grid capacity and the attendant Miller effect. But this is usually accompanied by an increase both in output and input capacities, so the improvement is not as great as might be expected. It has been said of capacity that two things may be done: 1) reduce it to the smallest possible value and 2) charge it. Additionally, the capacity may be distributed, but then the device is dynamic and its voltage memory is lost. In any case, good circuit dress and a high ratio of gm/C is indicated.

Inductive compensation of the circuit capacities allow a marked improvement in rise time, the amount being determined by the degree of overshoot that may be tolerated. The speed may be increased about 40% over an RC network if a 25% overshoot is allowed. This compensation will not relieve the Miller effect, so detrimental to triode operation. To do this in triodes requires some scheme of cross-neutralization. This is normally not very successful at high speeds.

Inasmuch as the initial slope of an exponential is the greatest, further improvement in transition speed may be accomplished by limiting plate and/or grid excursions. At least a two-to-one improvement in rise time may be accomplished if care is taken not to add appreciable capacity with the diodes.

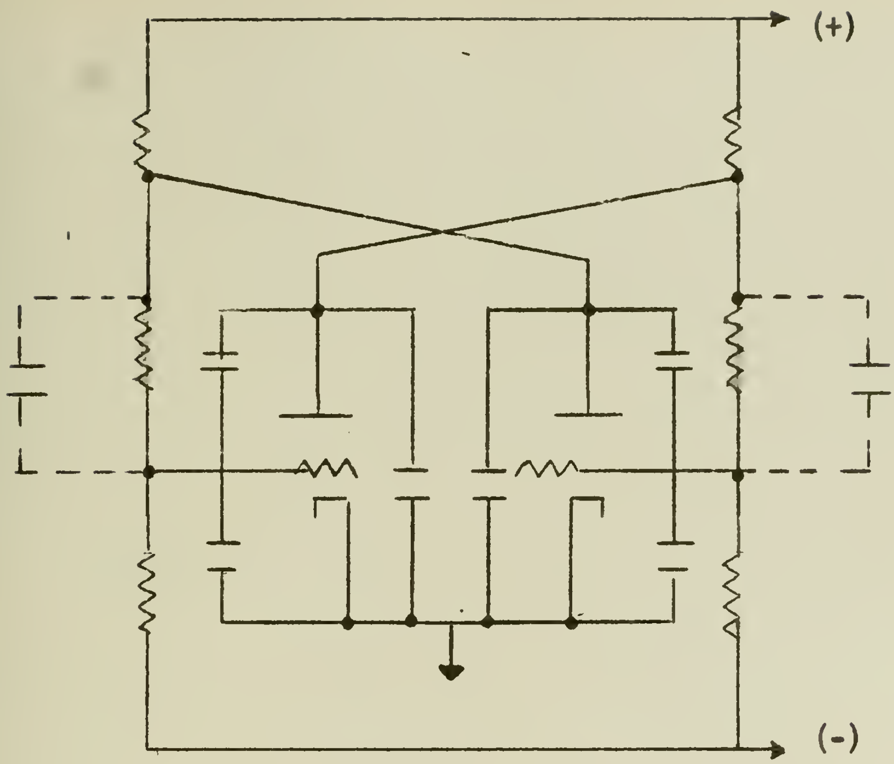


Figure 1. A Typical Eccles-Jordan Circuit

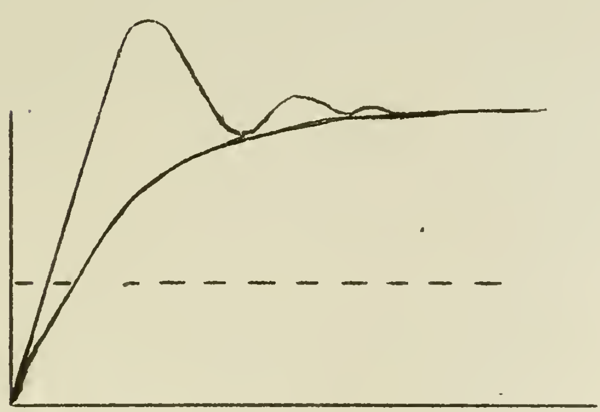


Figure 2. Influence of Compensation and Clamping

It is readily seen from the sketch in Figure 2 that combined clamping and inductive compensation offer an insignificant improvement over clamping alone. Considering the unavoidable increase in stray capacity and self-resonance of the inductance, the circuit will probably show performance inferior to that due to clamping alone.

Valley and Wallman [16] show, for rise time measured from 10% to 90% in a low-pass amplifier with small overshoot (less than three percent), that

$$(1) \qquad \frac{\text{gain}}{\text{rise time}} = \frac{G}{t_r} = \frac{gm}{2.2\,C} \quad \text{per microsecond,}$$

where, t_r is the rise time, G is the gain and gm is the transconductance. This relation allows a figure of merit for comparing tubes for pulse applications.

Taking as typical of present day tubes suitable for pulse applications the 6AH6, 6AK5, 6CB6 and the 6CL6 and assuming each is loaded with its own input and output capacity plus five mmf stray capacity, the figures of merit are 240, 228, 220 and 227, respectively. Average gm for these tubes is 9, 5, 6.2 and 11 ma/volt, respectively. This shows that, in general, higher transconductance is associated with higher input and output capacities, so the increase in figure of merit is not great.

Assuming further that a gain of unity is necessary to permit cascading of counter stages, and that clamping of upper and lower plate swing is used, an unclamped gain of approximately four will be needed. Assume also that the dead or settling time is equal to the rise time. There obtains a transition time of about 20 musec. This indicates a maximum counting rate of 50 mcs. Twenty-five mcs is more likely achievable.

The Hewlett-Packard Model 524A Frequency Counter [9] is an excellent example of the use of these techniques to produce a highly reliable 10 mcs counter. The Maintenance Manual prepared for this equipment gives a good practical treatment of the considerations pertinent to the design and operation of such systems.

A recent circuit development [8] employs clamping techniques with the Phillips EFP-60 secondary emission hexode (gm is 25 ma/volt). A pulse resolution time of 10 mμsec is obtained. Again assuming one half the pulse repetition period as rise time, and further that the equivalent bandwidth is one-half the reciprocal of rise time, we find

(2) $$\beta = \frac{1}{2t_r} = \frac{1}{10^{-8}} = 100 \text{ mcs}.$$

This is perhaps near the upper limit of bandwidth for lumped constant circuits.

4. Distributed Constant Circuits.

The bandwidth may be extended to about 500 mcs by using conventional tubes in a distributed amplifier system. When this is done, however, the system is dynamic, and frequency memory must be used. The operation is then of a double-side-band nature, requiring twice the bandwidth of the low-pass system. Conceivably a counting rate up to about 250 mcs could be achieved in this way. If further improvement is to be accomplished, the bandwidth of the Traveling Wave Amplifier is needed.

5. Frequency Memory Systems.

The need for increased bandwidth has led to the investigation of frequency memory for information storage and counting. The bandwidth available in dynamic systems may be used in this way to achieve much higher transition rates than are possible with lumped constant systems.

The principle work in this field has been done at the Laboratory for Electronics, of Stanford University, under a long term study contract with the Office of Naval Research. The work has been, logically, both of a theoretical and an experimental nature. Several reports have been issued on this work. Some of them are specifically referred to in this paper.

6. State of the Art.

The Stanford group has constructed several equipments in pursuit of this work. Principle among them are

1) an audio frequency system in the range two to four kcs. A magnetic tape was used to obtain a 1/8 second recirculation delay.

2) a fuzed quartz delay line recirculation system in the region of 15 mcs. Three-hundred fifty stable modes were obtained at one kilocycle intervals.

3) an eleven-mode lumped-constant system. A transitron oscillator was used as a decimal register.

4) a twelve-mode crystal-controlled oscillator in the region of six mcs. Satisfactory switching was obtained on a single-shot basis.

5) a bistable oscillator with normal modes at 200 and 257 kcs. Circuitry permitted keying from one frequency to another by a suitable input pulse.

The first two of these are four-terminal recirculation systems, while the remaining ones are of a two-terminal nature. All are only of general interest here, except the last one. The bistable oscillator will be described briefly in the next section.

7. Frequency Memory Binary Scaler.

This system [12] uses a standard transitronoscillator, except that two tank circuits are used in the screen lead. They are connected in series. One is tuned to each of the desired mode frequencies. It is known that this system cannot support two modes simultaneously, so oscillation will be sustained on the mode first excited, until changed by external influence. This circuit is the first practical counter using a two-state frequency memory.

A special balanced modulator is used for switching. When the inputs are supplied by the bistable oscillator and a local oscillator tuned to $f_o = f_1 + f_2$, where f_1 and f_2 are the mode frequencies, the modulator output is that of the non-oscillating mode. The signal so obtained is connected to the oscillator input through an amplifier and a vacuum tube switch. When the switch is closed, the gain of the oscillator at the operating frequency is reduced by the input signal, the oscillation dies out and the mode corresponding to the input signal is energized. Switching of modes is thus accomplished. The process is reversible with identical input pulses, so binary counting can be accomplished.

This system was reliably switched, but the switching rate was quite low. This is attributable to the energy storage in the tuned circuits, inasmuch as the energy in the oscillating circuit must be dissipated and a comparable storage achieved in the other to accomplish a transition. To a first approximation, the build-up or decay time is inversely proportional to the Q of the tuned circuits. From this it follows that considerations of stability and speed require compromise.

A distinct disadvantage of this system is the amount of auxiliary circuitry required. The additional equipment could likely be justified for a system of several modes intended for counting to a higher base, or as an information storage register. It could not be justified in a binary counter unless a substantial increase in speed could be achieved over an Eccles-Jordan circuit.

Difficulty of speed and stability associated with energy storage may be circumvented by use of distributed systems and active filters. This theme will be more fully developed in the remainder of the paper.

CHAPTER II

BASIC PRINCIPLES OF MULTI-MODE OSCILLATORY SYSTEMS

1. Oscillatory Systems.

Electronic systems capable of sustained oscillations must have one or more active elements to supply power lost in its passive parts and to a useful load. For the device to start, the initial power gain must be greater than that required to sustain oscillation, giving rise to a factor that grows with time. A gain reduction factor that is a function of amplitude is essential to reduce the loop gain as the signal builds up. This permits a stable amplitude for sustained oscillation to be reached. A satisfactory system must possess all of these features.

An excess of small signal gain necessary to sustain oscillation is easily obtained in most situations. The gain controlling nonlinearity is usually, though not always, associated with the active element of the system.

A frequency sensitive network, usually linear and passive, is necessary for mode determination. Networks of interest here are of the four-terminal type, involving delay. All distributed circuits involve delay, giving a factor $e^{-p\tau}$ in the Laplace transform of the transfer function, and are four-terminal networks.

In addition to mode determination, the multi-mode system must provide some form of mode discrimination if stability is to be achieved. In other words, some mechanism must operate to reduce the loop gain of all non-oscillatory modes to a value less than one. This mechanism must be a function of the degree of saturation of the desired mode. If this discrimination is not provided, simultaneous oscillation on two or more frequencies is possible, given a suitable system non-linearity. Alternatively, the system may shift rapidly from one mode to another. Experience has shown [6] that recirculation systems involving loop delay are particularly susceptable to this type of instability.

2. Linear Analysis of Four-Terminal Oscillators.

Linear analysis is useful to give insight into the nature of the oscillatory system. In particular, it shows the performance of the system from time zero until the amplitude of the oscillation reaches a value causing the system to become non-linear, assuming zero initial conditions. The response to any kind of excitation may be determined, assuming the excitation is mathematically expressible and that the resulting equations have a solution.

The essentials of linear analysis applicable to both two- and four-terminal systems have been reported by Lee [13]. Some of the more important features of his report, pertaining to four-terminal systems only, are reviewed below for convenience.

The analysis is based on a linear amplifier of gain μ having no delay, in series with a delay line having a magnitude factor β and a delay τ. The output of the delay line is connected to the input of the amplifier, closing the loop. This is shown schematically in Figure 1. It is assumed that μ and β are functions of frequency but not of time or signal magnitude.

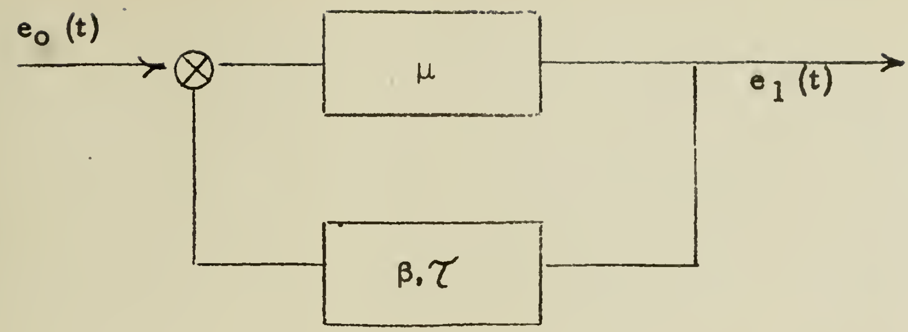


Figure 1. A Four-Terminal Delay-Type Recirculating System

Voltage equilibrium requires

(1) $\mu\, e_o(t) = e_1(t) - \mu\beta\, e_1(t - \tau)$

In the complex frequency plane, the transfer function is

(2) $$\frac{E_1(p)}{E_o(p)} = \frac{\mu}{1 - \mu\beta\, e^{-p\tau}}$$

For oscillation $E_o(p) \equiv 0$, $E_1(p) \neq 0$, giving

(3) $\mu\beta = e^{p\tau}$

which is the characteristic equation for the system. Values of p satisfying this equation are roots of the system and determine its natural modes. If these modes are independent they are called normal modes.

Solution of equation (3) is necessary to determine the roots. Taking natural logarithms,

(4) $(\sigma + j\omega)\tau = \ln|\mu\beta| + j[\phi + n(2\pi)]$

from which

(4a) $\sigma\tau = \ln|\mu\beta| = \alpha$

and

(4b) $\omega\tau = \phi + n(2\pi)$

where n is any integer. This gives two equations that must be satisfied simultaneously. There is no straight-forward solution here, since α, ϕ, μ and β are each functions of σ and ω. But in practice, $\sigma \ll \omega$ and $\phi \ll \omega$, allowing the reasonable approximation

(5) $\qquad \omega\tau = n\,(2\pi)$

and

(6) $\qquad \sigma = \dfrac{\ln|\mu\beta|}{\tau}$

This allows a complex frequency plot of σ versus ω when the forms of μ and β are known, with roots occurring for $\omega\tau$ equal to any multiple of 2π radians. Lee shows such a plot for β = a constant and

(7) $\qquad \mu = \dfrac{k}{1 + \frac{p}{\omega_o} + \frac{\omega_o}{p}}$

Roots occurring in the right half plane are active, while those occurring in the left half are dormant. The active roots are characterized by positive values of σ and represent signals that grow in time, the expansion rate being determined by the magnitude of σ.

Information on the manner in which the system amplitude will build in response to an input signal is gained from equation (2) by assuming a small sinusoidal signal of constant amplitude is applied at time zero. As may be noted from Figure 1., the input is not affected by the output until a time τ has elapsed: therefore, in the first interval τ the output is just

(8) $\qquad E_1\,(p) = \mu\,E_o(p), \; 0 < t < \tau.$

In the next interval the driving signal is reinforced by the output and

(9) $\qquad E_1\,(p) = \mu[\,E_o(p) + \mu\beta\,E_o(p)\,e^{-p\tau}]\; 0 < t < 2\tau.$

In the third interval the input is again reinforced and

$$(10) \qquad E_1(p) = \mu\,[E_o(p) + \mu\beta\, E_o(p)\, e^{-p\tau} + (\mu\beta)^2\, E_o(p)\, e^{-2p\tau}] \qquad 2\tau < t < 3\tau$$

while in the forth interval

$$(11) \qquad E_1(p) = \mu\,[E_o(p) + \mu\beta\, E_o(p)\, e^{-p\tau} + (\mu\beta)^2\, E_o(p)\, e^{-2p\tau} + (\mu\beta)^3\, E_o(p)\, e^{-3p\tau}] \qquad 3\tau < t < 4\tau$$

etc., allowing us to write, for the general interval,

$$(12) \qquad \frac{E_1(p)}{E_o(p)} = \mu\,[\,1 + \mu\beta\, e^{-p\tau} + (\mu\beta)^2\, e^{-2p\tau} + \cdots + (\mu\beta)^{(n-1)}\, e^{-(n-1)p\tau}]$$

$$= \sum_{1}^{n-1} [\mu\beta e^{-p\tau}]^{(n-1)} \qquad (n-1)\tau < t < n\tau$$

Inasmuch as

$$(13) \qquad e^{-p(t - m\tau)} = 0, \quad m\tau > t$$

the range restrictions may be dropped and equation (12) written simply as

$$(14) \qquad \frac{E_1(p)}{E_o(p)} = \mu \sum_{o}^{n} [\,\mu\beta\, e^{-p\tau}]^n$$

This expression is perfectly general, and can be used to represent any form of driving function that is Laplace transformable. In the case cited the output amplitude is seen to increase stepwise in time at a rate determined by the value of σ and the relation of the driving frequency to the natural modes of the system, the greatest rate obtaining when it coincides with a mode. Lee points out (op. cit.) that, if the driving function were a pulse of length τ of an exponential sinusoid, the output would be an exponentially growing wave with no discontinuities.

3. Mode Determination.

Possible modes of oscillation are determined by equation (4b), which requires

(4b) $\omega\tau = \phi + n(2\pi)$

If the phase constant is taken as zero for convenience,

(15) $\omega_1\tau = n_2\ (2\pi)$

and

(16) $\omega_2\tau = n_2\ (2\pi)$

Choose $n_2 = n_1 + 1$ so the modes are adjacent, and combine equations (15) and (16). There results

(17) $f_2 - f_1 = \delta f = \dfrac{n_2 - n_1}{\tau} = \dfrac{1}{\tau}$

Thus the modes are arithmetically spaced across the pass band of the amplifier. All modes having the same small signal gain are equally preferred, since they have the same expansion rate.

4. Mode Selection.

The bistable oscillator requires two modes of approximately equal preference. It is desirable that the small signal gain of all other modes be substantially below these. In general, this condition may be obtained in one of two ways:

1) Reduce loop delay. This separates two adjacent modes, such that one may fall near each half-power point in the small signal gain curve. All other modes occur essentially outside the pass band.

2) Use filters to remove undesired modes.

The mode transition rate is enhanced by small loop delay and large bandwidth, so these items are listed in order of preference. It follows that control of loop delay should be used to accomplish the desired mode selection where feasible. If this is impractical, filtering must be used.

Passive filters should not be employed to accomplish mode selection, because they function by limiting bandwidth. Because of the transient nature of the switching phenomenon, bandwidth imposes an upper limit on the transition rate. Indeed, passive filters cannot be used in some instances, because of space limitations. This suggests investigation of active filters of the feedback type, for they do not compromise the available bandwidth of the amplifier.

5. Multiple Feedback.

Let us modify the system of Figure 1 by adding a parallel feedback path, and examine the performance. The new system is shown in Figure 2. The equilibrium equation for this system

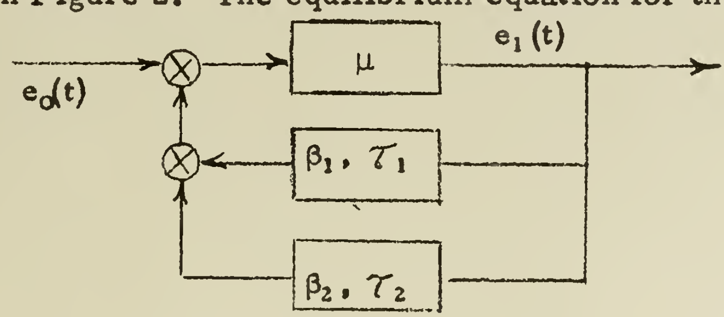


Figure 2. A Parallel Feedback System

may be written as

(18) $$e_1(t) = \mu e_o(t) + \mu\beta_1 e_1(t-\tau_1) + \mu\beta_2 e_1(t-\tau_2)$$

In the complex frequency plane,

(19) $$\frac{E_1(p)}{E_o(p)} = \frac{\mu}{1 - \mu\beta_1 e^{-p\tau_1} - \mu\beta_2 e^{-p\tau_2}}$$

When the transfer function is compared with that of equation (2)

$$(2) \qquad \frac{E_o(p)}{E_1(p)} = \frac{\mu}{1 - \mu\beta e^{-p\tau}}$$

we see that a term has been added to the denominator by the second feedback path. This term provides some new properties that may greatly modify the performance of the system.

The characteristic equation of the system is

$$(20) \qquad \mu\beta_1 e^{-p\tau_1} + \mu\beta_2 e^{-p\tau_2} = 1$$

System modes correspond to zeros of the characteristic equation. Again, there is no simple or exact solution, but modes occur when

$$(21) \qquad \mathrm{Re}[\beta_1 e^{-p\tau_1} + \beta_2 e^{-p\tau_2}] \geq \frac{1}{\mu}$$

and

$$(22) \qquad \mathrm{Im}[\beta_1 e^{-p\tau_1} + \beta_2 e^{-p\tau_2}] \equiv 0$$

The expansion rate of the modes is proportional to the excess of small signal gain. This, in turn, is proportional to the excess of equation (21). Mode selectivity may be accomplished by reducing the excess of equation (21) sufficiently for undesired modes, while maintaining or increasing it on desired ones. To see what relations need exist for this, we must examine the system parameters.

Equation (20) may be re-written as

$$(23) \qquad \beta_1 e^{-\sigma\tau_1} e^{-j\omega\tau_1} + \beta_2 e^{-\sigma\tau_2} e^{-j\omega\tau_2} \geq \frac{1}{\mu}$$

and the parameters are more readily recognized. For convenience, consider the parameters real, and define

$$(24) \qquad \beta_1 e^{-\sigma\tau_1} = A, \quad \beta_2 e^{-\sigma\tau_2} = B$$

Then equation (23) becomes

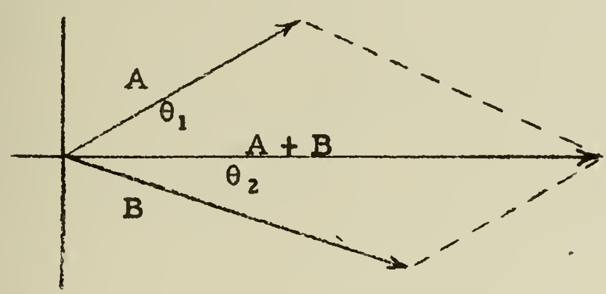
(25) $\quad A\ e^{-j\omega\tau_1} + B\ e^{-j\omega\tau_2} \geq \frac{1}{\mu}$

This seems to be the phasor sum of two quantities, one of magnitude A at an angle $-\omega\tau_1$ radians and one of magnitude B at an angle $-\omega\tau_2$ radians. It will be noted that they are both real and positive for $\omega = 0$.

Equation (25) is diagrammed in Figure 3. It is apparent that the largest excess occurs when both quantities are real and positive. This occurs when

(26) $\quad \omega\tau_1 = n_1(2\pi)$, and $\omega\tau_2 = n_2(2\pi)$

simultaneously.



$\theta_1 = n_1\ (2\pi) - \omega\tau_1$

$\theta_2 = n_2\ (2\pi) - \omega\tau_2$

Figure 3. A diagram of $A\ e^{-j\omega\tau_1} + B\ e^{-j\omega\tau_2}$

Solving each of equations (26) for frequency gives

(27) $\quad f = \frac{n_1}{\tau_1} = \frac{n_2}{\tau_2}$

From which

(28) $\quad \frac{\tau_1}{\tau_2} = \frac{n_1}{n_2}$

Adding equations (26) and solving for frequency yields another form.

(29) $\quad f = \frac{n_1 + n_2}{\tau_1 + \tau_2}$

It will be recognized that the n's are the electrical loop lengths, measured in wavelengths at frequency f.

We see from the foregoing relations that the individual terms of the characteristic equation will be real and positive each time its loop contains an integral number of wavelengths. Equation (23) will be real when the imaginary parts sum to zero. The manner in which this occurs is determined by the relation between the delay constants. In particular cases, the system may be examined graphically, or equation (25) may be converted to trignometric form and plotted. It will be noted that the phasor sum will be real, positive and a maximum simultaneously only if the component magnitudes are equal. Otherwise

(30) $\quad \mathrm{Im}\left[A\, e^{-p\tau_1} + B\, e^{-p\tau_2} \right]$

when the maximum occurs.

If we consider the special case in which $A = B$ and $\tau_1 = \tau_2$, we find equation (19) becomes

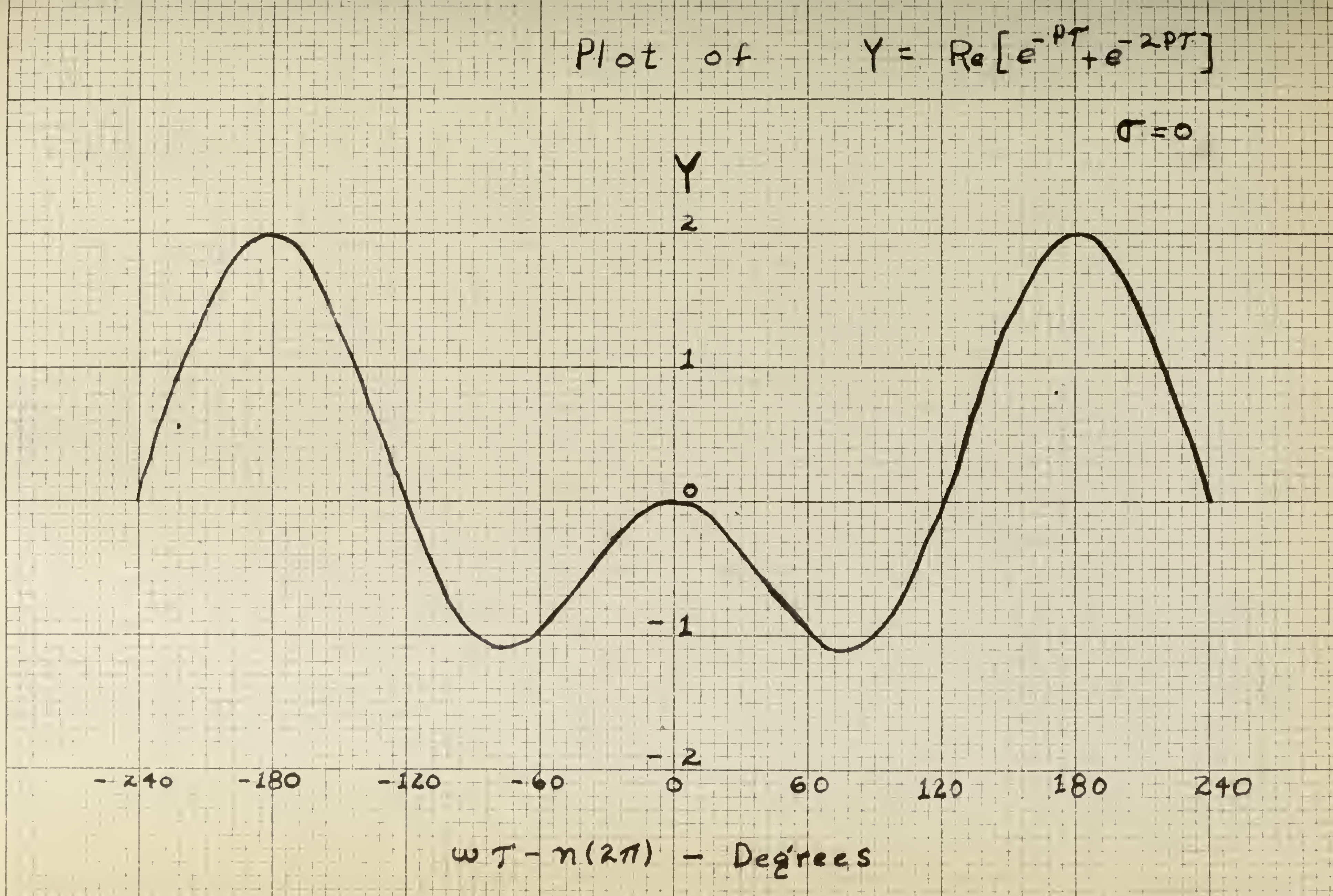
(31) $$\frac{E_1(p)}{E_o(p)} = \frac{\mu}{1 - 2\,\mu\beta e^{-p\tau}}$$

which displays the same characteristics as the system with the single feedback path.

In the case for which $2\tau_1 = \tau_2$ and $\beta_1 = \beta_2\, e^{-\sigma\tau}$ the magnitudes of the components are equal, but one component rotates twice as much for a change in frequency as does the other. The characteristic equation may then be written as

(32) $$e^{-j\omega\tau} + e^{-j2\omega\tau} \geq \frac{1}{A}$$

These terms are real and positive simultaneously each time the first is real and positive. The second is real and positive each time the first is real and negative. If the second term is considered as



Plot of $Y = Re\left[e^{-p\tau} + e^{-2p\tau}\right]$
$\sigma = 0$
Y
2
1
0
-1
-2
-240
-180
-120
-60
0
60
120
180
240
$\omega\tau - n(2\pi)$ - Degrees

generating the modes the first is seen to remove alternate modes and enhance the remaining ones. The variation of the real part of equation (32) is sketched in Figure 4.

It is noteworthy that filtering has been accomplished to the extent of removing half the modes, with no reduction in bandwidth, since the gain factor μ has not been influenced.

The principles outlined here may be applied to any feedback case. In instances where delay may be distributed with the gain, it is only necessary to note that there will be a delay between the input and output, and further, that the delay constants in the denominator of the transfer function include the total delay in the loops involved.

The transfer functions of several other common feedback systems will be found in appendix I at the end of this paper, together with some brief remarks relating to their properties.

6. Stability Considerations.

Linear analysis of a multi-mode oscillatory system will determine its potential oscillatory modes. If the system is to be useful, certain stability requirements must be met. Stability, in general, depends on the non-linear characteristics of the system. Treatment of the stability problem has been reported at some length by DeGrasse [4] for both two- and four-terminal systems. A brief summary of stability requirements relating to four-terminal systems, sufficient for the present purpose, will be given here.

A multi-mode oscillator must be multi-stable to be of practical value. Specifically, this means that once the oscillator is energized in a mode it must attain and maintain steady state operation at that frequency until changed by external influence. This implies the requirement that the energized mode become the preferred mode upon excitation. For this condition to prevail, an amplitude gain dependence is essential and it must operate on all possible modes.

As an illustration of the aforementioned requirements, consider an ideal system, of the delay type, in which the feedback factor is constant and the gain factor is a function of signal magnitude only. There would be infinitely many equally preferred modes, each having identical small signal gain. The modes would be selected in a random way if noise were allowed to excite the system at successive times. Stability in any selected mode would require that the small signal gain at all other modes be reduced such that their loop gains are less than one. Assuming an appropriate non-linearity, the system will oscillate simultaneously on all modes for which the small signal loop gain is greater than one.

Such a system would possess the characteristic that the oscillatory modes would be arithmetically spaced with $\delta f = 1/\tau$, where τ is the total loop delay. Each mode would be excited nearly equally upon turn-on, and many modes would have built up to appreciable amplitude before one of them began to limit gain by reaching the magnitude of non-linear operation. All these signals would combine in the (then) non-linear system to produce intermodulation products having the characteristics of both amplitude and phase modulation. The phase modulation would be of small deviation, producing a single pair of significant sidebands very similar to those of the amplitude modulation. Thus, the modulation products while differing in origin, are not distinguishable in their effects. Experience has shown that these sidebands are, in large measure, separated by $2\delta f$, and symmetrically placed about the strong mode: the correct circumstance for exciting further the two adjacent modes. This excitation could result in the growth of the adjacent modes more rapidly than could be accounted for by random excitation. The growth could, in fact, be sufficiently rapid that the adjacent modes could begin limiting the gain in the strong mode more rapidly than the strong mode limited that of adjacent modes, until one of the latter became dominant. This would result in a mode shift, and the whole operation could be repeated with a new mode playing the central role.

Edson has shown (op. cit.) that this condition could be avoided if the modes could be so chosen that no mode was the average of any pair in the system. This is, of course, not possible in the delay system. Each mode is the average of any two modes that are equally spaced therefrom.

In many cases, amplitude limiting (using diodes) will eliminate instability attributable to amplitude modulation; the transmission of small signals in the presence of a limited signal is reduced more rapidly than is that of the large signal for an increase in amplitude of the latter (Edson, op. cit.).

It is well known that amplitude limiting is desirable in phase- and frequency-modulation receivers, so it is evident that limiting will not remove this type of modulation. Stability in the presence of phase modulation has been achieved, however, by detection of the phase modulation in the output system, then phase modulating the input with the resulting signal in such a manner that these products are degenerate. The 350 mode oscillator mentioned in Chapter I was stabilized in this way.

As previously mentioned, considerations of stability in light of non-linear theory have been extensively reported by DeGrasse (op. cit.). A study of amplitude limiters has been reported at length by Amo [1]. The interested reader is referred to these works for further details.

CHAPTER III

THE TRAVELING WAVE AMPLIFIER AS AN OSCILLATOR

1. Basic Properties.

The traveling wave tube is essentially a helical transmission line arranged concentrically with an electron beam. Electromagnetic waves supported by the helix are propagated at slightly less than free space velocity along the wire length and at a velocity very much less than that of free space along the axis of the helix, depending on radius and pitch. When the electron beam voltage is correctly adjusted the electron velocity coincides with the phase velocity of the helix wave, and the two are in synchronism. Under these conditions there is coupling between the fields of the beam and those of the helix. Under appropriately adjusted conditions the ac wave on the beam and helix grow at the expense of dc beam energy, and amplification occurs.

The usefulness of the TWT arises from its distributed constant nature, allowing a very wide range of frequencies over which amplification can be reasonably uniform. A full octave range is not uncommon. While the power conversion efficiency of these devices is quite low by conventional standards, the bandwidth is very attractive. It is indeed this feature that makes it potentially valuable as a multimode oscillator.

The small signal (linear) gain of a TWA is approximately a linear function of helix length, on a decibel basis. The delay associated with the gain is also a linear function of helix length. Typical values for S-band tubes are 10 db and two millimicroseconds per inch, respectively.

Of great importance to satisfactory performance of the tube as an amplifier is the continuity of intrinsic wave impedance. This is so because the helix is a bi-directional transmission line, and any reflected energy due to impedance variations will be propagated as a backward wave, with a standing wave being the immediate consequence. Any secondary reflection of this backward wave closes a feedback loop with the usual results.

Quite a variety of methods have been developed for coupling a signal into and out of a helix, but the most popular one presently is a short length of contra-wound helix of appropriate dimensions, forming a directional coupler [2, 17]. This arrangement has the properties of an unbalanced transmission line, so may be coupled directly to a coaxial line. Fairly uniform coupling may be achieved over a band of frequencies comparable to the amplification band of the TWT.

Of importance later is the fact that the coupling of helix to beam is very loose. This means essentially that beam modulation cannot be removed. Thus, if the beam and helix passes through an output coupler, even though the helix wave is completely removed, the beam modulation is not appreciably affected, and will subsequently re-excite the helix wave.

To a first approximation, the gain of a TWA is proportional to the cube root of beam current, so amplitude modulation of a signal is possible. This is true, whether the signal is applied to the input or self-generated. In the latter case, however, the non-linearities associated with saturation effects must be taken into account.

The electrical length of the helix, for fixed frequency input, is proportional to beam velocity over a small range. Since beam velocity is proportional to helix voltage, phase modulation is readily achieved.

Saturation of the TWA is best considered in terms of power output for power input. As a function of power input, the power output increases linearly, declines from linearity, reaches a maximum, then falls off. This leads to the concept of "effective tube length". Maximum tube length may be thought to correspond to maximum power output; further increase in power input reduces the output and the effective tube length. If this is continued, the effective tube length approaches zero, and the power output becomes very small. We may profitably identify degrees of saturation with power gain above the linear range. This has been studied, both theoretically and experimentally for single-frequency input [15], but unfortunately, not for multiple-frequency input. No data is available on modulation products for two-signal input as a function of beam saturation. These data would be very useful in switching considerations.

2. The Oscillator.

Any system capable of power gain may sustain oscillations, and in this way the TWA is no exception. The fact that it is a distributed constant high gain device makes stabilizing it against oscillation for amplifier service a challenging problem indeed. Providing flat input and output coupling, over the wide frequency range over which the amplifier operates, is a near-impossible task; the output coupler is a difficult reflection source to suppress. The absence of cold-loss attenuation of the tube represents exposure to double jeopardy, for not only is the forward gain increased, but the attenuation of the backward wave is reduced. For this reason, attenuation is deliberately introduced into most amplifiers.

The amplifier is most readily converted into an oscillator by connecting a suitable fraction of the output back to the input. Selective filters may be inserted in the path if a single frequency or some particular group of frequencies are desired. This provides a wide range of possibilities, since only a relatively small fraction of the possible power output is needed to sustain the oscillation at a particular frequency. A very large number of potential modes can be obtained by increasing the delay in the feedback path, in view of the prevailing relation $\delta f = 1/\tau$ for mode separation. In this way $\tau\beta$ modes are theoretically attainable, where τ is the total loop delay and β is the available bandwidth.

The minimum delay attainable with external feedback represents a limitation, however, when a small number of modes is desired without artificial filtering. To circumvent this difficulty, one must use a shorter portion of the tube to reduce the delay (assuming, of course, that the feedback path has otherwise been shortened as much as possible). A feedback system may be placed on a reduced length of the tube in a variety of ways. Additional couplers may be installed between the normal input and output couplers; or, one or more couplers may be used in conjunction with the normal couplers. Connecting the output of one coupler to the input of a preceeding one, totally or fractionally, is one method of feed. Perhaps the simplest way, when the resulting standing waves can be tolerated, is to deliberately introduce an impedance discontinuity at an appropriate point on the helix. A suitable discontinuity to produce a secondary reflection will assure oscillation, providing only that there is sufficient gain between the two points of reflection.

A limit will be reached in reducing the number of modes in this way when the length of tube helix provides the minimum gain that will support oscillation. In one case tried the loop delay was reduced to 2.5 mμsec, giving $\delta f \approx 400$ mcs, with adequate gain. Further reduction in modes suggests consideration of multiple loop techniques.

3. Mode Selectivity.

Further reduction of possible modes after loop delay has been minimized requires filtering of some nature to suppress undesired modes to the extent that they will not build up in preference to desired modes. Passive filters introduce an inordinate degree of difficulty in view of the fact that they must be installed within any encapsulation of the tube. Active filters of the multiple feedback variety apparently offer possibilities that have not been thoroughly investigated. The basic theory of multiple feedback systems is worthy of considering in any case, for it is not likely, it appears at this writing, that all inadvertent feedback paths can be adequately suppressed to prevent their affecting the performance of the system. Even admitting the possibility, it may well be more convenient to "join 'em" than to elect the alternative.

As an illustration of possible feedback paths, desired or otherwise, consider Figure 1. This represents an amplifier on which has been placed one additional coupler, with one intentional loop. Recalling that the couplers are not perfect and that the unterminated high level end of the helix may extend appreciably beyond the last coupler, it is found that there are essentially four cascaded amplifiers and eleven possible feedback loops. These lead to the formidable block diagram of Figure 2. It is obvious that prediction of circuit performance would be quite tedious at best.

It should be noted in passing that the circuit alluded to above is more easily realized than not. Also, it is capable of sustaining oscillations on several frequencies simultaneously. Limited access to the circuit for measurement make it very difficult to deduce what the conditions in the circuit actually are.

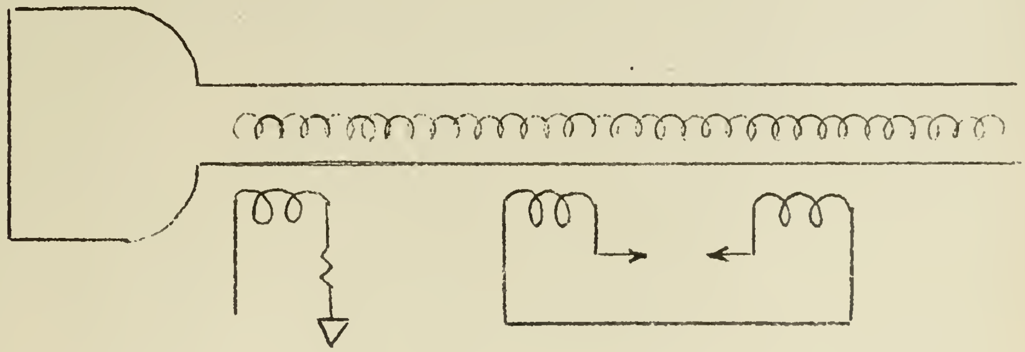


Figure 1. A Typical Oscillator Circuit

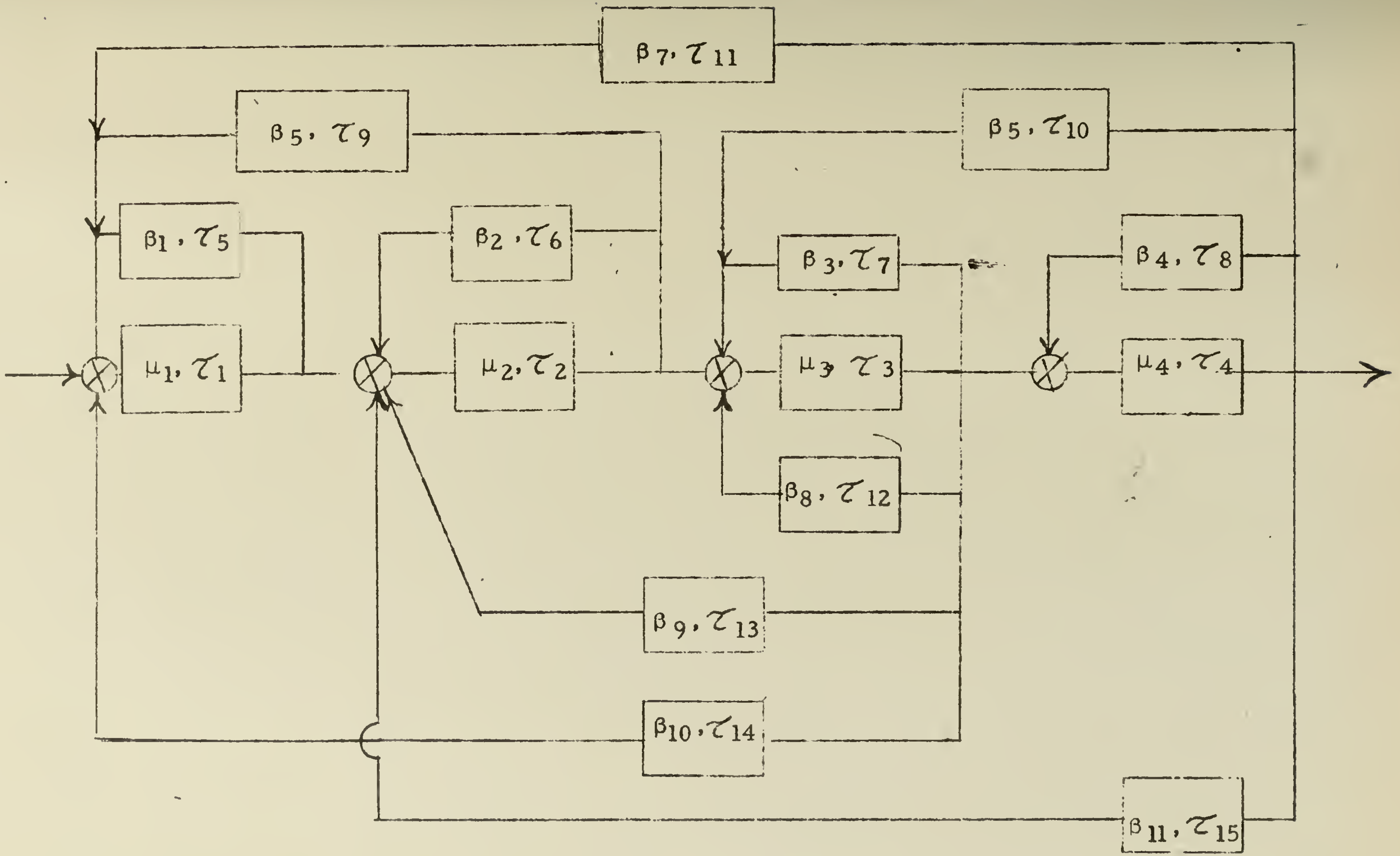


Figure 2. A Block Diagram of Figure 2.

Assuming that undesired feedback can be suppressed or controlled, multiple feedback loops appear to be potentially quite useful in suppressing undesired modes while simultaneously increasing the preference of the desired ones. The use of such selection may be particularly valuable, as indicated previously, in cases where insideous regeneration may be caused to perform a useful function.

4. Mode Stability.

The statement made earlier that mode stability was a prerequisite to obtaining a useful system applies fully to the traveling wave oscillator. The nature of the problem does not differ substantially from that in the more conventional system.

If the system uses external feedback (external in the structural sense) the stabilizing methods developed by Edson, et al, are applicable without modification, within limits. Reduction of phase modulation should prove to be simpler than in the conventional low frequency system, because of the inherent ability of the tube to produce phase modulation by variation of helix voltage. It would only be necessary to apply phase detection to the output, then apply the resulting signal to the helix in proper phase.

When the feedback loop includes only a portion of the tube, the problem becomes somewhat more difficult, largely from a mechanical point of view. Electrical considerations may differ from those of the external system in some cases.

It would be difficult or impossible to use passive amplitude limiters due to space limitations, in all cases of internal feedback. It is, or course, impossible to use these when a backward helix wave is employed to close the feedback path. Filtering methods, probably of the feedback type, will need be used. It may be possible with feedback techniques to stagger the modes, thus providing a measure of stability. Mode stability will depend on long-term average of helix voltage and beam current. Excellent regulation, particularly of helix voltage, will be essential.

5. Mode Detection.

Determination of the mode that is in operation is necessary. This is the information that makes multi-mode operation desirable.

Mode detection is normally a problem of frequency determination and is discussed by Disman [5]. This may be accomplished by a variety of methods varying in degree of complexity. The method chosen would depend on the requirements of the application.

When the number of modes is small, say four or less, matters can usually be arranged such that the magnitudes of oscillation are discrete. In this case, mode detection is resolved to the relatively simple problem of amplitude discrimination of the rectified output. This information can then be properly processed to actuate other equipment.

6. Mode Switching.

Achieving mode stability assures that once a mode is excited it and it alone will remain excited until acted upon by external influence. This affords the quality of memory. If the memory feature is to be useful, it is necessary that the system respond to the appropriate external signals in a systematic way. It must shift in some sequence through the set of operating modes in response to a given number of identical signals, so that counting is accomplished. The shift of operation by one mode through the operating sequence in response to an input signal is referred to as mode switching, or, simply as switching. A mode shift not in response to an input signal is an instability.

Disman [5] discusses the general requirements of switching. The remainder of this discussion will be concerned with the switching of modes in a bistable traveling wave oscillator, employing internal loops.

Fundamentally, switching consists of (1) disabling the oscillator and (2) exciting the desired mode. The oscillation is stopped in any manner that allows the small signal loop gain to be reduced to a value less than one, while mode excitation requires a signal of appropriate frequency and of magnitude several times greater than that produced by noise at undesired mode frequencies.

The TWO may be arranged so that it can be represented as two or more cascaded gain sections, with one or more of these enclosed in a feedback loop, to produce oscillation over some portion of the tube in such a way that there is a gain section ahead of the oscillating loop. This affords an opportunity of modulating the electron beam, by means of a suitable input signal, so that the degree and the nature of beam saturation, upon the entry into the oscillatory loop, is controllable; in this way, the loop gain is controllable. One possible arrangement is shown schematically in Fig. 3. With no input signal the loop will sustain oscillation in a mode determined by circuit parameters. If

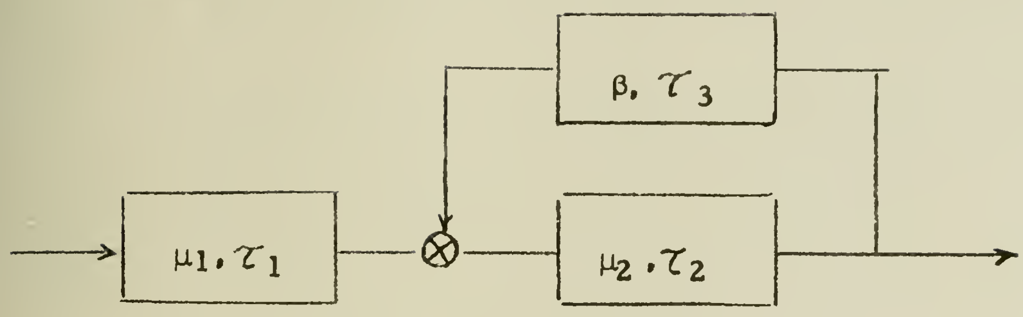


Figure 3. A Block Diagram

now, a signal of such magnitude as to cause beam saturation at the output of μ_1 is injected, the gain μ_2 will be reduced to, or near, zero at the mode frequency. In this way the oscillation will be stopped. If the injected signal corresponds to a mode frequency, the mode will be excited at saturation amplitude. When the injected signal is removed, the output of μ_2, at saturation level, will be impressed on a "clean" beam at the input, through the feedback loop, for a period $\tau_2 + \tau_3$ before the self-excitation level can be changed. The magnitude of the oscillation will then adjust itself for the requisite degree of beam saturation at the output of μ_2, and mode switching will have been accomplished.

Suppose an rf pulse of length $(\tau_2 + \tau_3) < \tau < 2\,(\tau_2 + \tau_3)$ is injected into the oscillating loop with the pulse frequency $f_o = f_1 + f_2$. The heterodyne products will be impressed on a "clean" beam at the input of the loop following the pulse. If the coefficient of the difference frequency, produced in the heterodyne process, is larger than that of the operating frequency, it is reasonable to expect mode switching to occur. If the procedure is repeated, a reversal may be expected, and binary switching would be accomplished with a single frequency input.

The frequency components resulting form a signal

(1) $E_i = V_m \text{ Cos } mt + V_n \text{ Cos } nt$

and a transfer function

(2) $E_o = a\, E_i + b\, E_i^2 + c\, E_i^3 + d\, E_i^4 + e\, E_i^5$

are shown in appendix III. The coefficient of the difference frequency component is

(3) $A = b\, V_m\, V_n + \frac{3}{2}\, d\, V_m^3\, V_n$

Let V_m be the coefficient of f_o (equal to one for convenience) and that of f_1 be V_n. There results

(4) $A = b\, V_n + \frac{3}{2} d\, V_n = V_n \left[\, b + \frac{3}{2}\, d \,\right]$

Obviously

(5) $\frac{A}{V_n} > 1$ if $2\,b + 3\,d > 2$

It follows that conditions described in the paragraph above may be achieved from physically realizable non-linearities.

As mentioned previously, the TWT has the property that the phase of the output relative to the input is a linear function of the length of helix employed and of helix voltage, for small variation of the latter. This makes phase modulation a simple matter and affords another possibility for mode switching.

The components of a phase modulated signal of unity amplitude are expressed as

$$(15) \qquad E_o = J_o(\Delta\phi) \sin \omega t$$
$$+ J_1(\Delta\phi) \left[\mathrm{Cos}\ 2\pi\ (f + \delta f)\ t + \mathrm{Cos}\ 2\pi\ (f - \delta f)\ t \right]$$
$$+ \text{other terms of higher order Bessel coefficients.}$$

where $\Delta\phi$ is the peak phase deviation and δf is frequency of phase variation. If $\delta f = 1/\tau$, where τ is the total loop delay, then δf is the mode separation of the oscillator. The primary sidebands generated correspond to modes adjacent to the operating mode. When the upper mode is operating the lower sideband is the useful one, and vice versa. For the bistable case, the presence of the two sidebands instead of only the desired one should cause no trouble.

The sideband amplitudes will be a maximum when $J_1(\Delta\phi)$ is a maximum, this occurring for $\Delta\phi \cong 2.0$; the corresponding value for $J_o(\Delta\phi)$ is approximately 0.340. But the strongest tendency to switch may occur when $J_o(\Delta\phi) = 0$, and this occurs for $\Delta\phi \cong 2.5$, with the corresponding value of $J_1(\Delta\phi) \cong 0.495$. It should be noted in passing that this situation corresponds to small-deviation modulation, and all sideband pairs higher than the second are insignificant.

This modulation can be achieved by modulating the helix voltage with a single cycle of a sine wave of frequency δf, which, of course, has a period τ, the total loop delay. An approximation of this pulse may be generated with an improperly terminated coaxial pulse-forming line. Difficulty may be encountered in driving the helix in this manner; for as a load, it will appear as an unterminated single-wire transmission line with a relatively large characteristic impedance, so reflections may be expected.

The products due to the transient nature of this type signal may be expected to differ considerably from those due to steady state phase modulation, on which equation (15) is based. The transient problem has not here been examined.

CHAPTER IV

LABORATORY PROJECT

1. Experimental Objectives.

A laboratory investigation was conducted to determine the feasibility of obtaining bistable oscillation of the TWA by employing a short regenerative loop. This study was intended to determine the degree of saturation necessary to satisfactory oscillation, the shortest length of tube over which this degree of saturation could be achieved, the type of instabilities to be confronted and means of mode selection.

Contingent upon successful accomplishment of bistable oscillation, it was desired to investigate mode switching by injecting a suitable signal into the normal amplifier input, as well as by pulsing the helix.

2. Equipment.

An -hp- 491A one-watt S-band amplifier was modified to permit pulse modulation of the beam current and thus the gain. In this way measurement of an otherwise CW signal was possible using standard 1000 cycle amplifiers. Alternately, the oscillations could be stopped briefly with a short pulse, then be allowed to build up again while the results were viewed on an oscilloscope. In addition to this, the capsule in which the tube is mounted was modified in various ways, as may be seen in what follows.

The equipment is briefly described in appendix II; the work performed is described in detail in appendix IV.

3. Measurements.

The TWA is a prolific oscillator, so obtaining oscillation did not constitute a difficulty per se. Achieving an oscillation of the desired nature did prove difficult. This was particularly true since it was, in general, not possible to make measurements directly on the circuits of interest. Observations were largely confined to the input and output of the system. The conditions of interest were of necessity, deduced therefrom. Since a component of any and all beam modulation appears in the output, it was difficult or impossible to ascertain quantities of interest.

4. Spurious Regeneration.

The principle difficulty encountered in the practical work was attributable to undesired regenerations. The impedance match between couplers and helix apparently depended on signal level (beam saturation), with reflections increasing at high levels. Backward waves on the helix, due to these reflections, encountered impedance discontinuities supplying secondary reflections to close a feedback loop. In many instances, these loops were strong enough to support oscillation. In one such instance, twenty-six frequencies were measured in the output.

A substantial percentage of power in the helix wave was found to pass through the couplers, or the helix wave was re-excited by the beam, the modulation of which is essentially unaffected by the couplers because of loose coupling. The results were quite the same in either case, as far as power leakage past a coupler is concerned. This power must be absorbed or reflections will occur to contribute regeneration.

Use of attenuators to reduce gain and/or absorb undesired power is not an easy or immediate solution; their presence caused a discontinuity in impedance for the helix wave, unless great care was taken to prevent it. This mis-match was shown capable of causing an oscillation.

5. Mode Stability.

In spite of (or perhaps because of) the spurious regenerations, bistable operation was obtained. The front section of the tube was attenuated to the point where a signal transfer loss of about five decibels was realized, after the oscillatory loop was moved to the extreme output end of the tube. Under these conditions, strong reflections were realized from the unterminated end of the tube helix: this had a strong influence on the oscillatory circuit. Bistable operation was obtained under five different combinations of beam current and helix voltage. These conditions were found to be critical, but not severely so. Some conditions were more critical than others.

These frequencies were obtained in pairs of adjacent modes from 1700 to 2710 mcs, with frequency separation of about 122 mcs, corresponding to a loop delay of about eight mμsec. It was found that the oscillations were very stable in mode, shifting only in response to external influence.

6. Mode Switching.

Difficulty was experienced when attempts were made to pulse the system with an rf signal at the sum frequency ($f_o = f_1 + f_2$) because of excessive attenuation on the front gain section and an unfortunate combination of mode frequencies. Because of these circumstances, a pulse of eight to ten mμsec duration at about a one watt level, and of frequency in the vicinity of 4400 mcs, was needed. The microwave switch available would only balance in the range of 2200 to 3200 mcs and the 4400 mcs frequency was nearly out of range for the one-watt amplifier available.

The system was pulsed at the mean of the mode frequencies, with reliance on harmonic generation in the beam to produce the sum frequency. Erratic switching was obtained, but this is an unsatisfactory way of keying the system.

Another method tried also proved to be unsuccessful. A small amplifier was used on the pulse out of the switch to drive a crystal harmonic generator followed by a waveguide high pass filter. The signal so obtained was passed through a C-band amplifier, which was used to drive a one-watt S-band amplifier. The pulse fidelity was apparently maintained and the amplitude was attained, but the accumulated noise, when impressed on the input of oscillator, destroyed mode stability.

Attempts to modulate the helix with a ten mμsec approximation of a sine wave, to produce phase modulation, were unsuccessful.

7. Conclusions.

This method of employing a TWA is promising. Switching rates upward of 100 mcs are achievable with S-band tubes, and perhaps higher with K- or X-band tubes. If the requisite stability against spurious regeneration can be attained, with a satisfactory margin of gain in the lead section of the tube, switching should be handily accomplished.

Further study of the application of multiple loop feedback theory to this system should prove quite beneficial. It appears feasible to dimension the system in such a way that regeneration that would otherwise be spurious and insidious could be caused to serve the useful purpose of providing some additional mode selectivity.

Finally, a tube could be constructed that has a helix just long enough to provide the necessary gain. It should be possible to use the tube in such a way that the unterminated ends of the helix furnished reflections to close the feedback loop. A single coupler would determine which mode was oscillating. For switching, the beam could be modulated prior to entering this loop. No attenuation would be needed and loop delay could be in the order of two to five mμsec.

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APPENDIX I

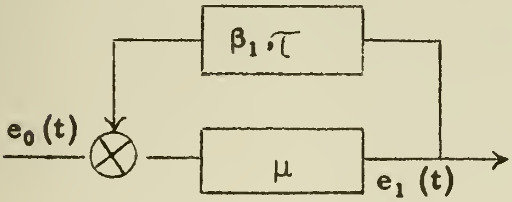
REGENERATION IN CERTAIN FEEDBACK SYSTEMS

1. General Considerations.

The considerations are here limited to four-terminal networks, and the re-circulation point of view is adopted. The gain and feedback factors are considered as real constants, although it is known that in an actual oscillating system the gain factors are functions of the complex variable P and of the non-linear gain characteristic on which all oscillators depend for amplitude stability; the feedback factor will be in practice a function of P to some degree. However, in a delay-controlled system, particularly with a wide band distributed-constant amplifier, the total loop phase is the principle element of frequency determination.

The complex notation is employed. The complex-frequency element σ is considered zero where convenient, since it is known that the dominant pole lies on the $j\omega$ axis in conditions of stable oscillation.

2. The Simple Single Loop.



(1) $e_1(t) = \mu[e_0(t) + \beta e_1(t-\tau)]$

(2) $E_1(p) = \mu[E_0(p) + \mu\beta E_1(p) e^{-p\tau}]$

Figure 1. A Simple Feedback Loop.

(3) $$\frac{E_1(p)}{E_0(p)} = \frac{\mu}{1 - \beta\mu e^{-p\tau}}$$

For sustained oscillations, $E(p) \equiv o$, $o < E_1(p) < \infty$

This requires

(4) $\mu\beta e^{-p\tau} = 1$, or $e^{p\tau} = \mu\beta$

Under the stated assumption this may be re-written as

(5) $\cos\omega\tau = \mu\beta = 1, \quad \sin\omega\tau = 0$

The necessary conditions to satisfy these relations are

(6) $\omega\tau = n(2\pi)$, n = any integer.

From this it is seen that

(7) $f_1 = \dfrac{n_1}{\tau}, \quad f_2 = \dfrac{n_2}{\tau}$

and

(8) $f_2 - f_1 = \dfrac{n_2 - n_1}{\tau}$

If $n_2 = n_1 + 1$, f_2 is the next higher frequency adjacent to f_1, and

(9) $\delta f = \dfrac{1}{\tau}$

Thus the regenerative frequencies are spaced **inversely** as τ and are ordered arithmetically for fixed τ. Each is equally probable as an oscillating mode, since theoretically each has infinite small signal gain.

3. Two Simple Loops Cascaded.

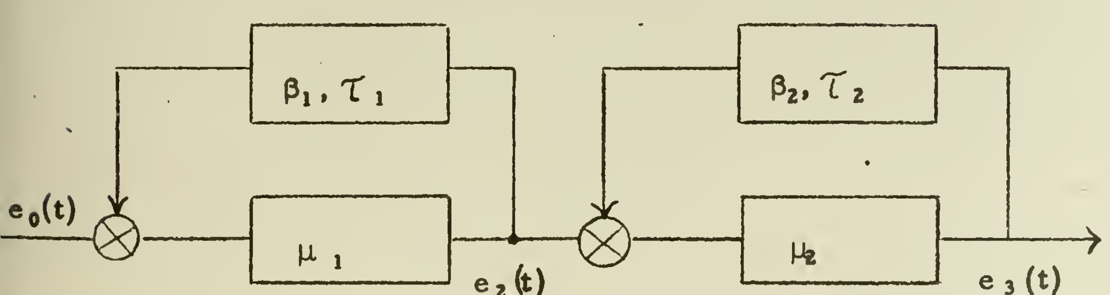


Figure 2. Two Simple Loops Cascaded

With the aid of relations from Section 2, this may be re-drawn as

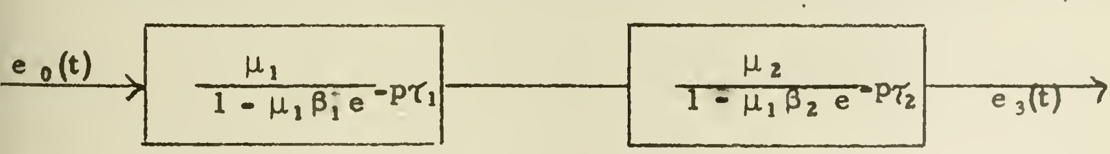


Figure 3. Equivalent Block Diagram

and the desired ratio written immediately as

$$(10) \quad \frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}]}$$

By previous reasoning, the conditions for sustained oscillations require:

$$(11) \quad [1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}] = 0$$

This condition is satisfied for

$$(12) \quad 1 - \mu_1 \beta_1 e^{-p\tau_1} = 0 \text{ or } 1 - \mu_2 \beta_2 e^{-p\tau_2} = 0$$

They may, of course, be zero simultaneously. This will occur when $\tau_1/\tau_2 = n_1/n_2$, where n_1 and n_2 are integers.

If the gain is reduced to a value too small to support oscillation the transfer function will go through maxima and minima with respect to frequency, as equation (11) goes through minima and maxima, respectively. It is sufficient here to consider the real part of equation (11) to determine the order of occurance of maxima and minima of equation (10). A sketch of this function for $\tau_2/\tau_1 = 4/3$ is shown in Figure 5, and for $\tau_2/\tau_1 = 5/3$ is shown in Figure 6. It should be noted that all maxima and minima will be of the same height if $\tau_1 = \tau_2$.

4. Three Simple Loops Cascaded.

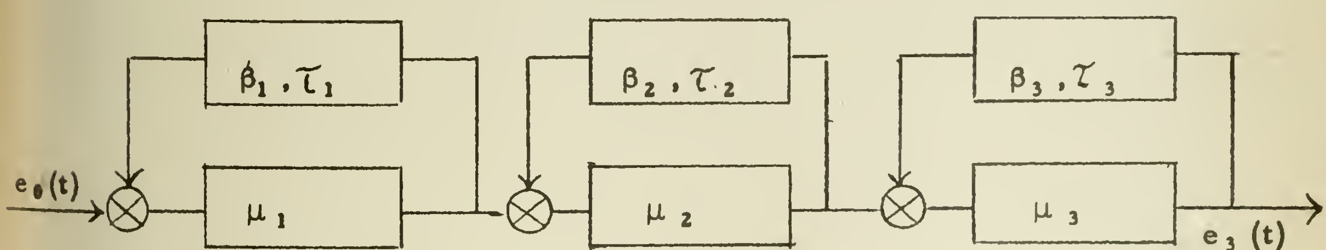


Figure 4. Three Cascaded Simple Loops

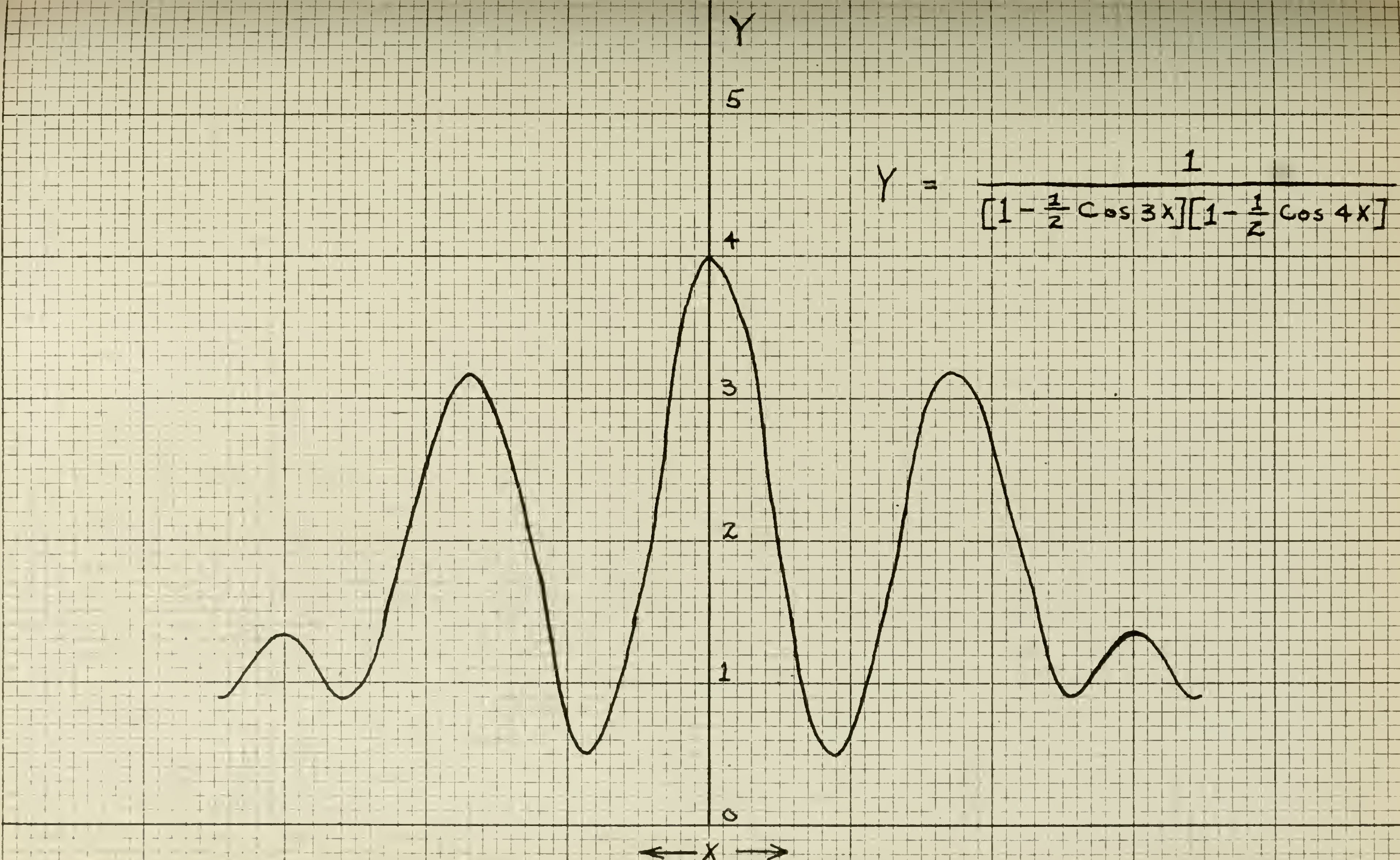


Figure 5

Appendix V

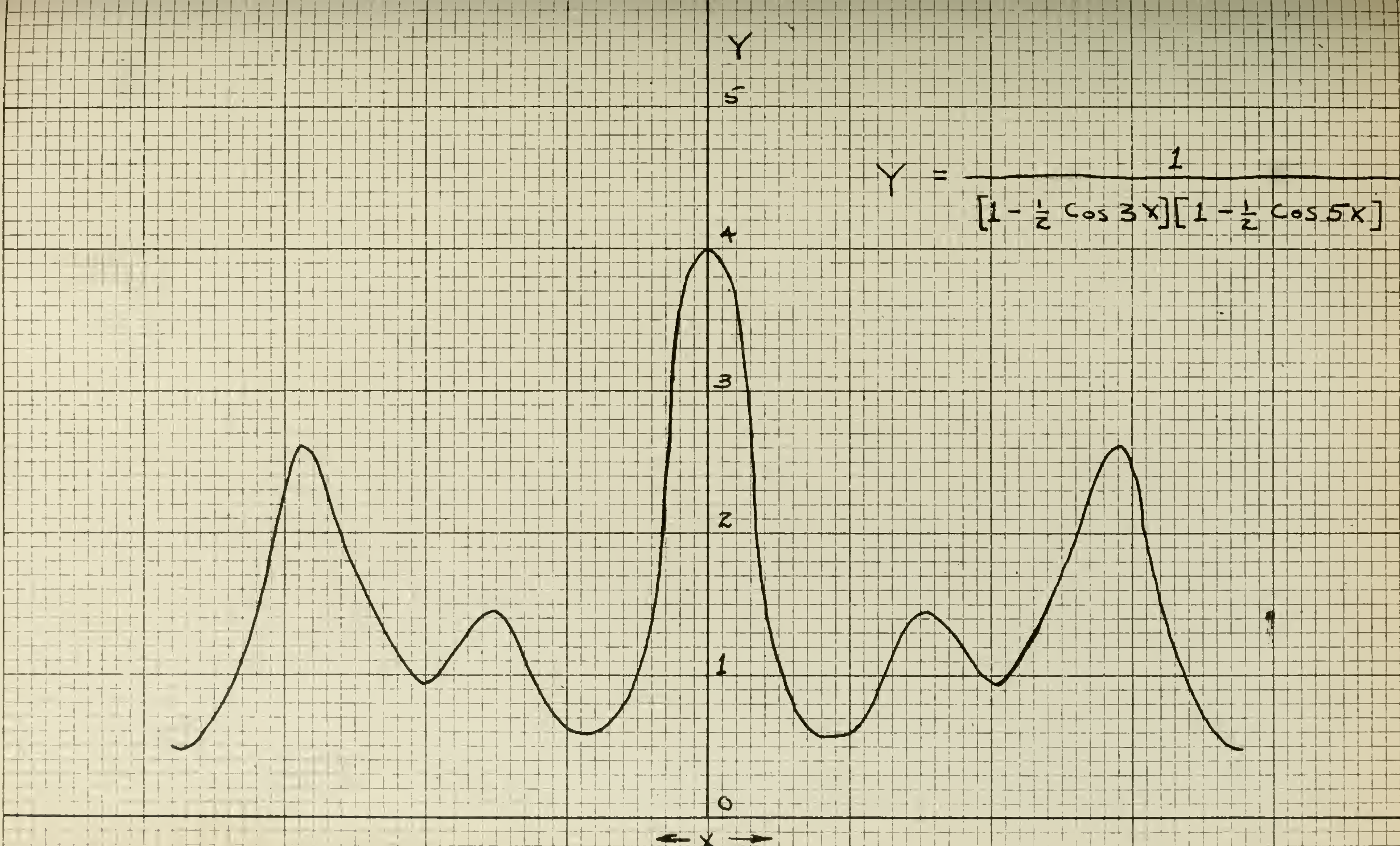


Figure 6

Appendix V

By reasoning from sections 2 and 3 the desired ratio may be written directly as

$$(13)\quad \frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2 \mu_3}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}][1 - \mu_3 \beta_3 e^{-p\tau_3}]}$$

This is essentially similar to the cascade of section 2, differing by an additional factor, and affording the possibility of three loops regenerating at once. Again, for finite gain the maxima and minima will be equal if $\tau_1 = \tau_2 = \tau_3$ and unequal otherwise. A sketch of this function is shown in Figure 8.

The required conditions for simultaneous regeneration are

$$(14)\quad \frac{\tau_1}{\tau_2} = \frac{n_1}{n_2}, \qquad \frac{\tau_2}{\tau_3} = \frac{n_2}{n_3}, \qquad \frac{\tau_3}{\tau_1} = \frac{n_3}{n_1}$$

where the n's have their previous significance.

5. Two Simple Loops Cascaded with an Overall Loop.

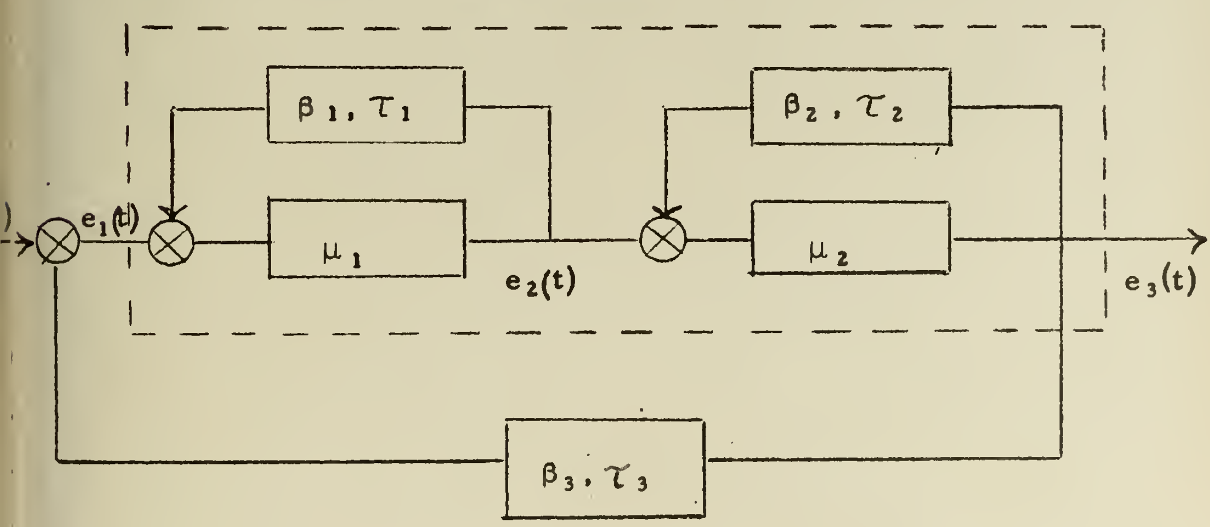
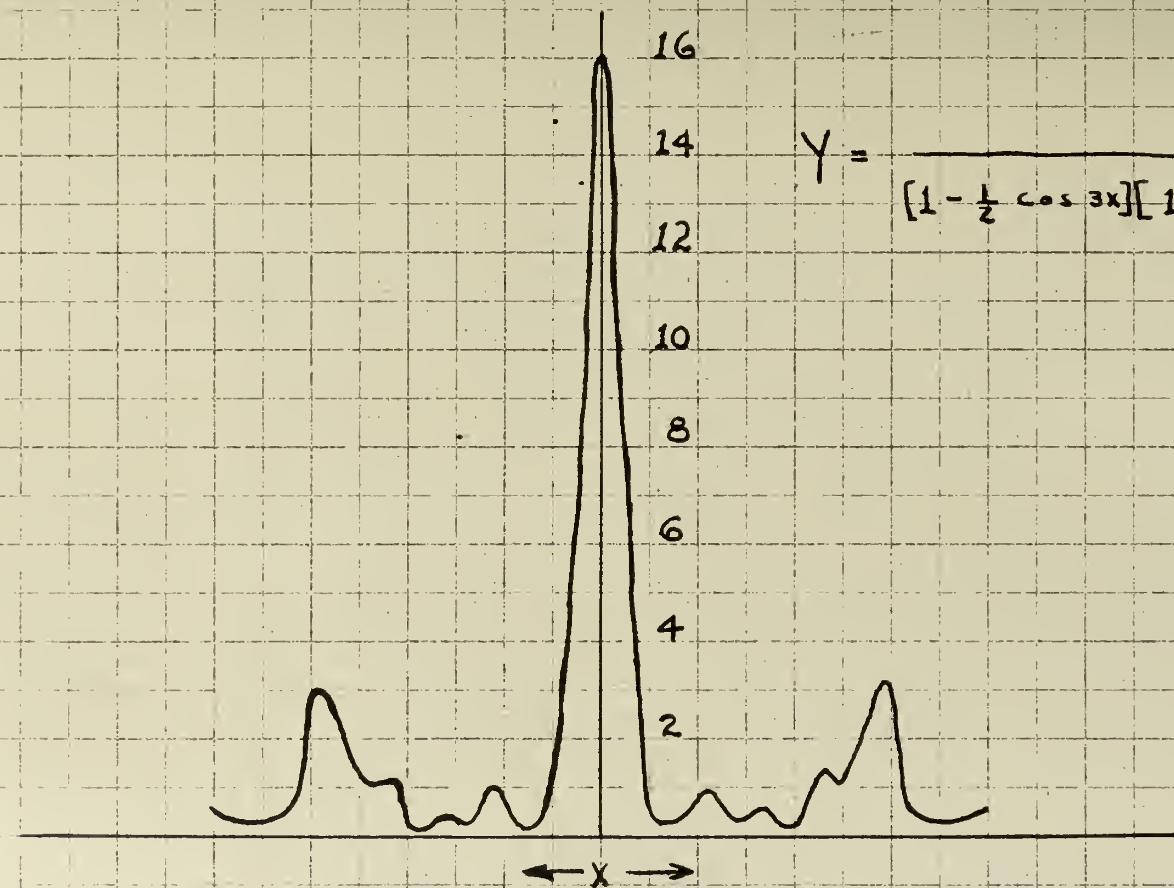


Figure 7. Two Simple Loops within a Major Loop



$$Y = \frac{1}{\left[1 - \frac{1}{2}\cos 3x\right]\left[1 - \frac{1}{2}\cos 4x\right]\left[1 - \frac{1}{2}\cos 5x\right]}$$

Figure 8

Appendix V

The equation for the enclosed portion of the system is, from Section 2,

$$(15) \quad E_3(p) = \frac{\mu_1 \mu_2 \; E_1(p)}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}]}$$

and the overall equation becomes

$$(16) \quad E_3(p) = \frac{\mu_1 \mu_2 [E_0(p) + \beta_3 E_3(p) \; e^{-p\tau_3}]}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}]}$$

from which we obtain

$$(17) \quad \frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}][1 - \mu_1 \mu_2 \beta_3 e^{-p\tau_3}]}$$

In the case of interest, $\tau_3 \cong \tau_1 + \tau_2$, and equation (15) is written

$$(18) \quad \frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2}{[1 - \mu_1 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}][1 - \mu_1 \mu_2 \beta_3 e^{-p(\tau_1 + \tau_2)}]}$$

Simultaneous regeneration of all three loops is obtained when

$$(19) \quad \omega\tau_1 = n_1(2\pi), \quad \omega\tau_2 = n_2(2\pi)$$

for then

$$(20) \quad \omega\tau_3 = \omega(\tau_1 + \tau_2) = (n_1 + n_2)(2\pi)$$

It is to be noted that these loops are not independent because of the $\mu_1 \mu_2$ product in the overall loop. Desirable selectivity characteristics may be inherently present, but circuit proportions are likely to be critical.

6. Major Loop and Simple Minor Loop

This is a simplification of the system of Section 4, and the transfer function is obtained from equations (16) by removing the appropriate factor from the denominator and noting that there is no inherent relation between the time constants.

The transfer function is

(21) $$\frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2}{[1 - \mu_1 \mu_2 \beta_1 e^{-p\tau_1}][1 - \mu_2 \beta_2 e^{-p\tau_2}]}$$

Simultaneous regeneration will be obtained when

(22) $$\omega \tau_1 = n_1 (2\pi), \quad \omega \tau_2 = n_2 (2\pi)$$

It is again noted that the loops are not independent.

7. Two Overlapping Loops.

This system includes two feedback loops with one common gain section, as shown schematically in Figure 9.

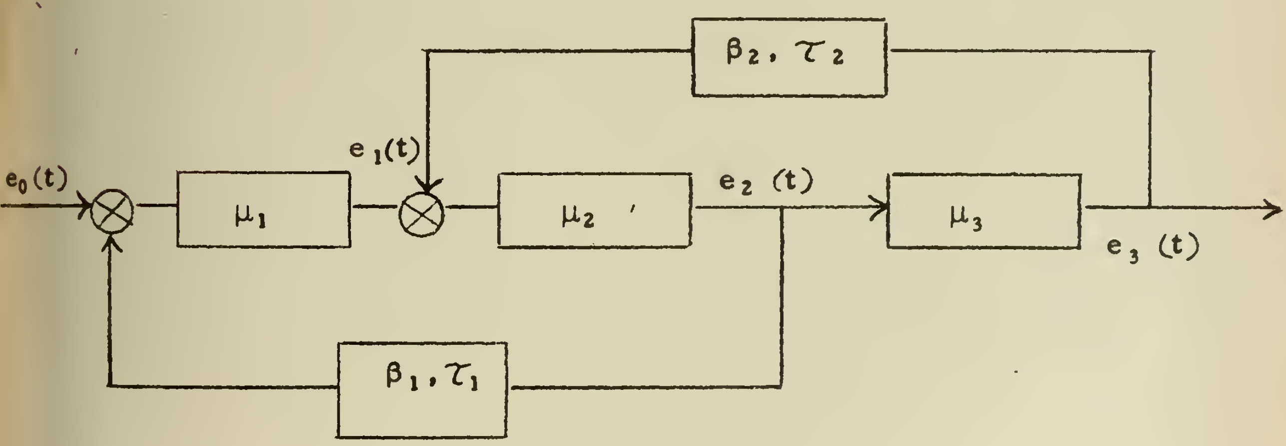


Figure 9. Two Overlapping Feedback Loops

Writing the equation for the system,

(23) $$E_1(p) = \mu_1 E_0(p) + \mu_1 \beta_1 E_2(p) e^{-p\tau_1}$$

(24) $$E_2(p) = \mu_2 \beta_2 E_3(p) + \mu_2 E_1(p)$$

(25) $$E_3(p) = \mu_3 E_2(p)$$

These are manipulated for the transfer function

(26) $$\frac{E_3(p)}{E_0(p)} = \frac{\mu_1 \mu_2 \mu_3}{[1 - \mu_1 \mu_2 \beta_1 e^{-p\tau_1} - \mu_2 \mu_3 \beta_2 e^{-p\tau_2}]}$$

Because of the element common to both loops the denominator cannot be zero by the influence of one loop independently of the other, as in previous cases.

For oscillation to be possible it is necessary that

(27) $\mu_1 \mu_2 \beta_1 e^{-p\tau_1} + \mu_2 \mu_3 \beta_2 e^{-p\tau_2} \geq 1$

This system possesses a very wide range of properties, depending on the relations of the coefficients and the loop delays. In the special case when

(28) $\mu_1 \mu_2 \beta_1 = \mu_2 \mu_3 \beta_2 = a$

it is necessary for

(29) $\omega\tau_1 - 2\pi n_1 = -[\omega\tau_2 - n_2(2\pi)]$

for equation (25) to be real. This condition is expressed by

(30) $f = \dfrac{n_1 + n_2}{\tau_1 + \tau_2}$

When the real portion of equation (25) is simultaneously larger than one, an oscillatory mode is obtained. A plot of the variation of the real portion of equation (25) will show potential modes and mode preference.

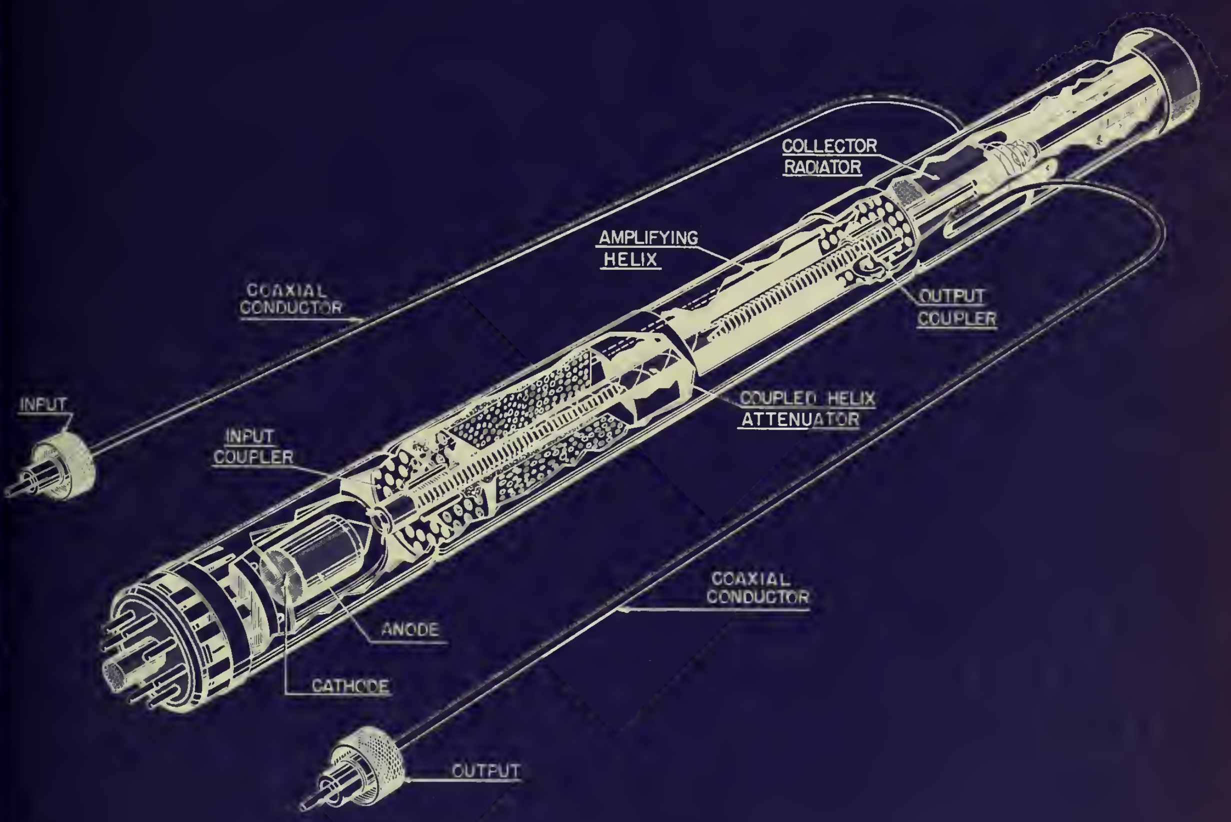
APPEXDIX II

EXPERIMENTAL EQUIPMENT

The instrument used as the experimental vehicle was a Hewlett-Packard Traveling-Wave Amplifier, Model 491A, serial 124. This instrument operates at S-band with a nominal range of 2 - 4 kilomegacycles, 30 decibels of power gain with a nominal maximum of one watt sinusoidal power output. The Huggins Laboratories HA-2 traveling-wave tube employing external helix input and output couplers is used. The equipment is described in detail in the Instruction Manual [10], and further data are available in a Hewlett-Packard Journal [11].

The heart of the system is the TWT amplifier capsule. This consists of a brass tube 1-15/16 inches in diameter and approximately 17 inches in length into which is fitted the input and output helical couplers, the TWT, necessary attenuators, tube sockets, collector radiators, collector leads, etc. A cut-away view of the assembled capsule is shown in Figure 1.

In order to allow reasonable access to the capsule when installed in the electromagnet, it is installed with the tube base toward the front panel. The capsule is fitted with an octal plug, so that it fits into a socket much as a vacuum tube would. Additionally, the gun structure of the TWT is designed to be operated immersed in a uniform longitudinal magnetic field, as is also necessary for the drift space of the tube. This made it desirable to bring out the coaxial leads from the input and output couplers at the collector end of the capsule, i.e., at the back of the instrument, where the geometry of



Model 491A Traveling-Wave Tube Amplifier Assembly
Cut-away View

Figure 1

II - 2

the tube allows the coaxial lines to be within the capsule until outside the uniform magnetic field. This requires the line from the input coupler to pass through the flanges of the output coupler, as can be seen from Figure 1. This led to some mechanical difficulty in our work.

The couplers used are contra-wound, co-directional helical couplers. They consist in helices of appropriate pitch and wire diameter impregnated with teflon dielectric, made close-fitting to the TWT drift tube for maximum coupling to the inner helix. The coupling helix is fitted into a sheath and spider designed to support the TWT. The sheath assembly forms a ground surface for the helix and for terminating the outer coaxial conductor. The dimensions are so arranged that the phase velocity on the coupling helix is appropriately matched to that of the inner (TWT) helix such that the nominal intrinsic wave impedance is 51 ohms, permitting an impedance match in the coaxial to helix transformation. Normal termination is provided to absorb any power not transferred to the inner helix to reduce reflections and thereby improve the VSWR.

The Huggins HA-2 TWT has a nominal power gain of approximately 10 decibels per inch of helix. There are slightly more than six inches of helix between the input and output couplers, indicating a potential power gain of about 60 decibels. Any attempt to use this much gain leads to instability due to the inability to isolate the input from the output of the amplifier sufficiently. This coupling can come about in three principle ways; 1) reflections due to mis-match between the TWT helix and the output coupler or by the output load being improper, 2) radiation of the signal from the high level end of the tube, thereby propagating a coaxial mode within the capsule and 3) at high beam currents collector secondary emission can produce a backward beam that travels through the helix with essentially full energy, thus allowing gain in the backward direction.

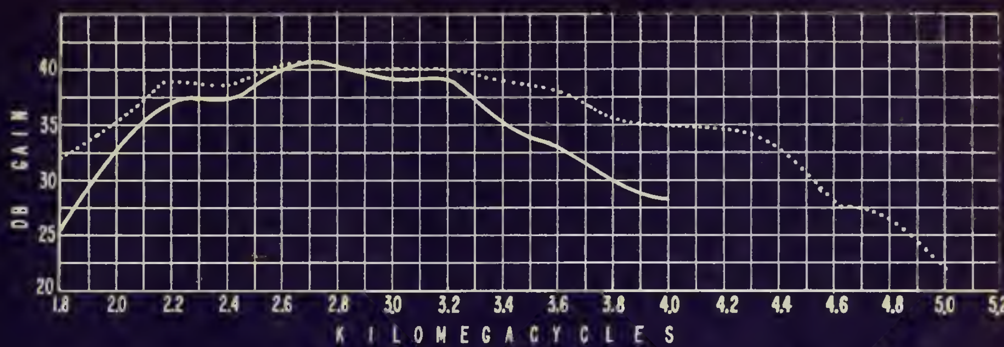
Reflections due to mismatch at the output coupler would be propagated backward down the TWT helix with essentially no attenuation other than that offered by cold-loss on the tube. The input coupler is an output coupler for the backward wave, and if the input load can absorb the associated power no instability would result, but any mismatch will reflect a fraction of the power in the forward direction and regeneration occurs. Power reflected due to improper load sees the output coupler as an input coupler and a backward helix wave is launched with the same results outlined. Stability against slow backward wave is obtained by sufficient attenuation in the form of a bifilar helix of lossy wire that couples equally to the backward wave and the forward wave. In addition, a suitable length of teflon sleeve with an aquadag liner which attenuates all slow waves equally well (for a given frequency) is used to assure stability when the output VSWR is maintained below 2:1.

The same considerations outlined above apply to stabilizing against backward beam amplification; but the most certain remedy is reduction of beam current and helix voltage.

Fast wave propagation mentioned in case three is attenuated by use of a porous dielectric block impregnated with powdered brass. It is about two inches in length and its diameter is the same as the inside of the capsule. It immediately follows the input coupler. The center hole through which the tube passes is sufficiently large that the slow wave is not effected.

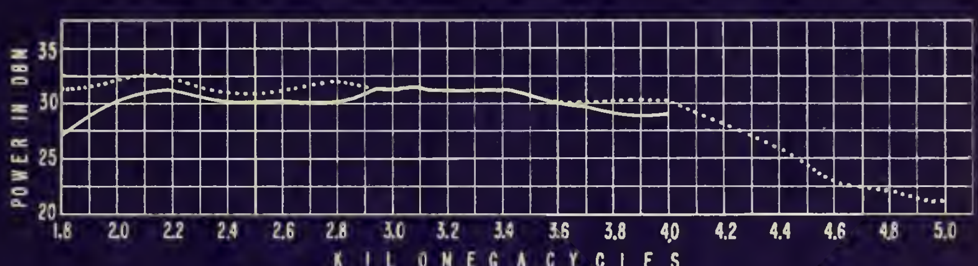
As previously mentioned, the power gain is a nominal 30 decibels with small signal operation; it exceeds this value by as much as 10 decibels for small signal input, and is at least equal to this figure over the normal frequency range with one milliwatt input. It is entering into saturation, however, and the maximum power out is slightly greater than one watt. The performance curves are shown in Figure 2.

The auxiliary circuits consist of the electromagnet and its power supply and a regulated source for each the anode and helix, together with appropriate metering and safety circuits. The anode supply regulator was modified to permit square-wave modulation of the beam current so that measurements of CW oscillations could be made with a tuned (1000 cycles) standing wave ratio indicator. The diagram of the normal circuit is given by Figure 3 and the modification is shown in Figure 4.



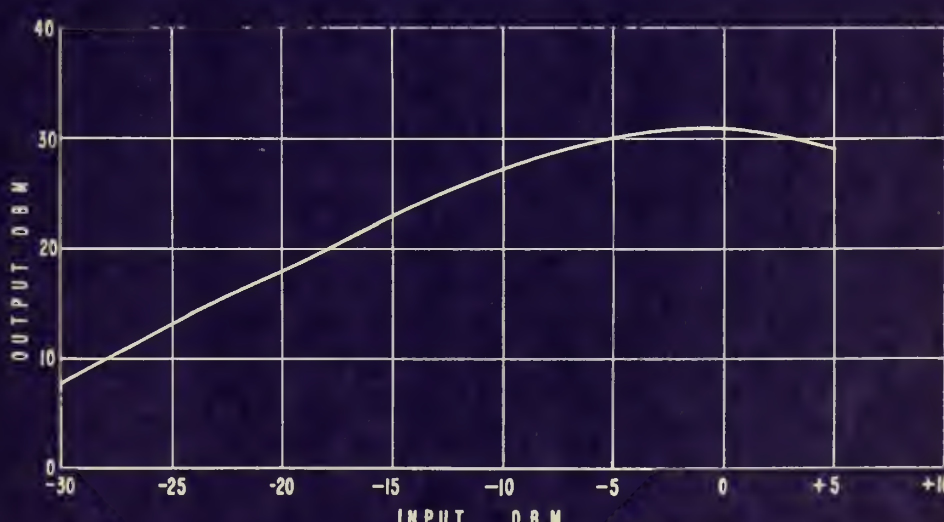
GAIN WITH +5 DBM OUTPUT

OPTIMUM HELIX VOLTAGE (dotted line)
FIXED HELIX VOLTAGE (solid line)

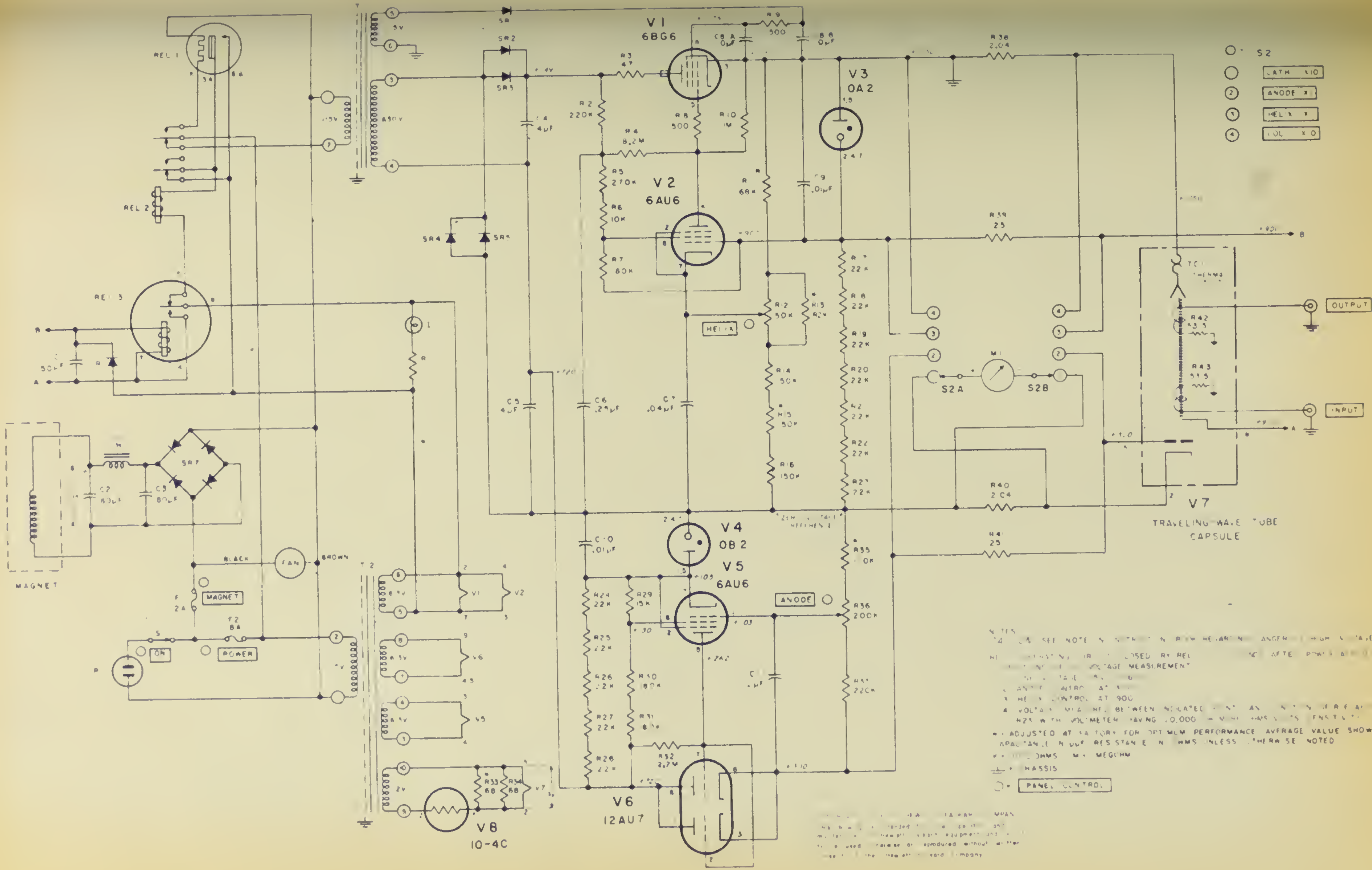


POWER OUTPUT WITH O DBM INPUT

OPTIMUM HELIX VOLTAGE (dotted line)
FIXED HELIX VOLTAGE (solid line)



INPUT POWER VS. OUTPUT POWER
AT 4 KMC



SCHEMATIC DIAGRAM OF MODEL 491A

SERIAL 69 & ABOVE

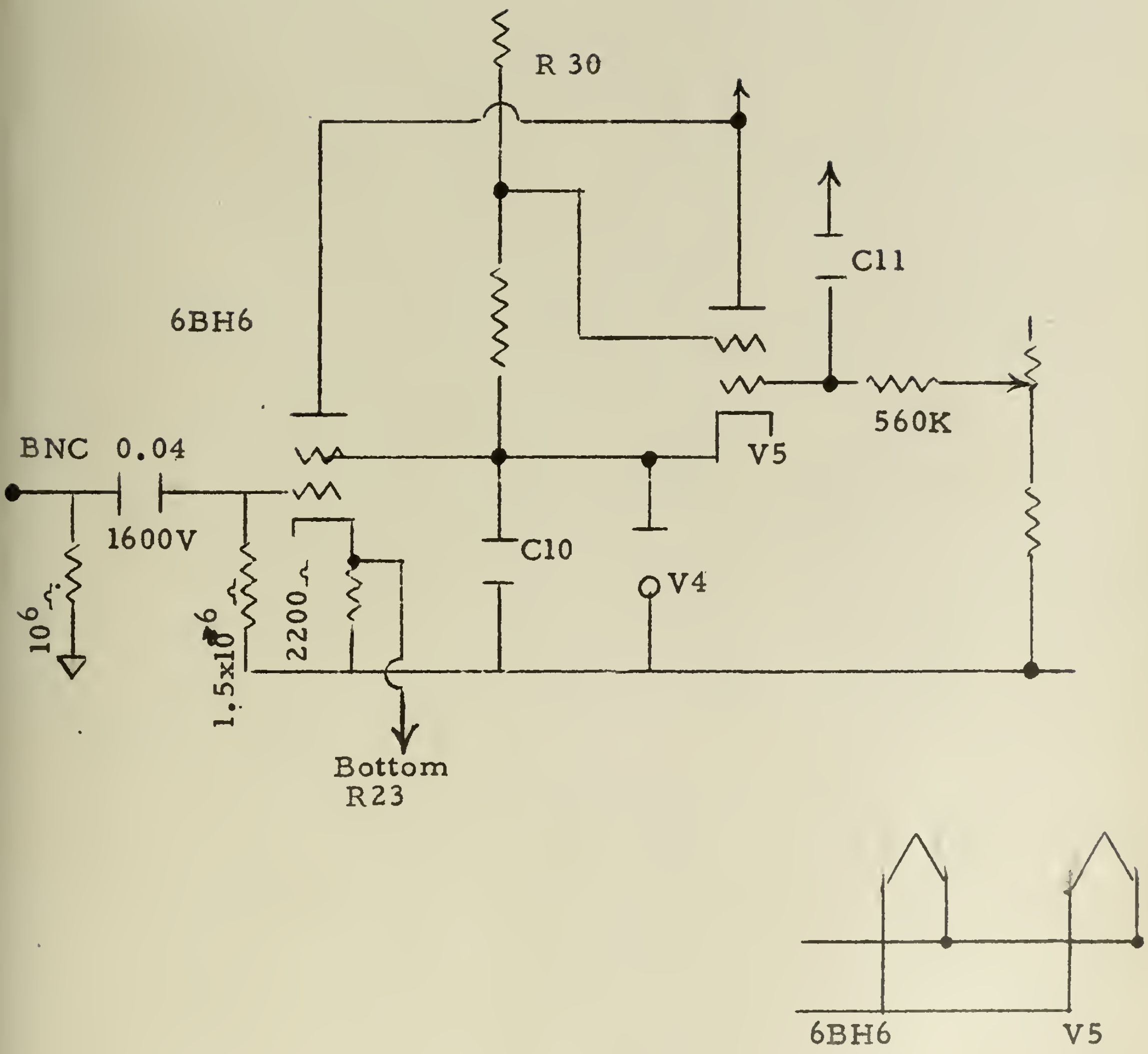


Figure 4. Beam Current Modulator

APPENDIX III

EXPANSION OF A FIFTH DEGREE TRANSFER FUNCTION WITH ARBITRARY COEFFICIENTS FOR TWO-SIGNAL INPUT

Consider the transfer function

$$E_o = a\,E_i + b\,E_i^2 + c\,E_i^3 + d\,E_i^4 + e\,E_i^5$$

with an input

$$E_i = V_m \text{ Cos } mt + V_n \text{ Cos } nt$$

and the output components are found to be

$$\begin{aligned}
E_o &= a\,(V_m \text{Cos } mt + V_n \text{ Cos } nt)\\
&\quad + b\,(V_m^2 \text{ Cos}^2 mt + 2\,V_m\,V_n \text{ Cos } mt \text{ Cos } nt + V_n^2 \text{ Cos}^2 nt)\\
&\quad + c\,(V_m^3 \text{ Cos}^3 mt + 3\,V_m^2\,V_n \text{ Cos}^2 mt \text{ Cos } nt\\
&\qquad + 3\,V_m\,V_n^2 \text{ Cos } mt \text{ Cos}^2 nt + V_n^3 \text{ Cos}^3 nt)\\
&\quad + d\,(V_m^4 \text{ Cos}^4 mt + 4\,V_m^3\,V_n \text{ Cos}^3 mt \text{ Cos } nt\\
&\qquad + 6\,V_m^2\,V_n^2 \text{ Cos}^2 mt \text{ Cos}^2 nt\\
&\qquad + 4\,V_m\,V_n^3 \text{ Cos } mt \text{ Cos}^3 nt + V_n^4 \text{ Cos}^4 nt)\\
&\quad + e\,(V_m^5 \text{ Cos}^5 mt + 5\,V_m^4\,V_n \text{ Cos}^4 mt \text{ Cos } nt\\
&\qquad + 10\,V_m^3\,V_n^2 \text{ Cos}^3 mt \text{ Cos}^2 nt + 10\,V_m^2\,V_n^3 \text{ Cos}^2 mt \text{ Cos}^3 nt\\
&\qquad + 5\,V_m\,V_n^4 \text{ Cos } mt \text{ Cos}^4 nt + V_n^5 \text{ Cos}^5 nt)
\end{aligned}$$

Collecting terms:

$$\begin{aligned}
E_o &= a\,V_m \text{ Cos } mt + a\,V_n \text{ Cos } nt\\
&\quad + b\,V_m^2 \text{ Cos}^2 mt + b\,V_n \text{ Cos}^2 nt\\
&\quad + c\,V_m^3 \text{ Cos}^3 mt + c\,V_n^3 \text{ Cos}^3 nt\\
&\quad + d\,V_m^4 \text{ Cos}^4 mt + d\,V_n^4 \text{ Cos}^4 nt\\
&\quad + e\,V_m^5 \text{ Cos}^5 mt + e\,V_n^5 \text{ Cos}^5 nt
\end{aligned}$$

Collecting terms (Cont'd.)

$$+ 2\,b\,V_m V_n \operatorname{Cos} mt \operatorname{Cos} nt$$
$$+ 6\,d\,V_m^2 V_n^2 \operatorname{Cos}^2 mt \operatorname{Cos}^2 nt$$
$$+ 3\,c\,V_m^2 V_n \operatorname{Cos}^2 mt \operatorname{Cos} nt$$
$$+ 3\,c\,V_m V_n^2 \operatorname{Cos} mt \operatorname{Cos}^2 nt$$
$$+ 4\,d\,V_m^3 V_n \operatorname{Cos}^3 mt \operatorname{Cos} nt$$
$$+ 4\,d\,V_m V_n^3 \operatorname{Cos} mt \operatorname{Cos}^3 nt$$
$$+ 5\,e\,V_m^4 V_n \operatorname{Cos}^4 mt \operatorname{Cos} nt$$
$$+ 5\,e\,V_m V_n^4 \operatorname{Cos} mt \operatorname{Cos}^4 nt$$
$$+ 10\,e\,V_m^3 V_n^2 \operatorname{Cos}^3 mt \operatorname{Cos}^2 nt$$
$$+ 10\,e\,V_m^2 V_n^3 \operatorname{Cos}^2 mt \operatorname{Cos}^3 nt$$

Expanding the powers of Cosines:

$$E_o = a\,V_m \operatorname{Cos} mt + a\,V_n \operatorname{Cos} nt$$
$$+ 1/2\,b\,V_m^2 + 1/2\,b\,V_m^2 \operatorname{Cos} 2mt + 1/2\,b\,V_n^2 + 1/2\,6\,V_n^2 \operatorname{Cos} 2nt$$
$$+ 3/4\,c\,V_m^3 \operatorname{Cos} mt + 1/4\,c\,V_m^3 \operatorname{Cos} 3mt$$
$$+ 3/4\,c\,V_n^3 \operatorname{Cos} nt + 1/4\,c\,V_n^3 \operatorname{Cos} 3nt$$
$$+ 3/8\,d\,V_m^4 + 1/2\,d\,V_m^4 \operatorname{Cos} 2mt + 1/8\,d\,V_m^4 \operatorname{Cos} 4mt$$
$$+ 3/8\,d\,V_n^4 + 1/2\,d\,V_n^4 \operatorname{Cos} 2nt + 1/8\,d\,V_n^4 \operatorname{Cos} 4nt$$
$$+ 1/16\,e\,V_m^5 \operatorname{Cos} 5mt + 5/16\,e\,V_m^5 \operatorname{Cos} 3mt + 5/8\,V_m^5 \operatorname{Cos} mt$$
$$+ 1/16\,e\,V_n^5 \operatorname{Cos} 5nt + 5/16\,e\,V_n^5 \operatorname{Cos} 3nt + 5/8\,V_n^5 \operatorname{Cos} nt +$$
$$+ b\,V_m V_n \operatorname{Cos}(m+n)t + b\,V_m V_n \operatorname{Cos}(m-n)t$$
$$+ 3/2\,d\,V_m^2 V_n^2 + 3/4\,V_m^2 V_n^2 \operatorname{Cos} 2(m+n)t$$
$$+ 3/4\,V_m^2 V_n^2 \operatorname{Cos} 2(m-n)t$$
$$+ 3/2\,d\,V_m^2 V_n^2 \operatorname{Cos} 2mt + 3/2\,d\,V_m^2 V_n^2 \operatorname{Cos} 2nt$$
$$+ 3/2\,c\,V_m^2 V_n + 3/4\,c\,V_m^2 V_n \operatorname{Cos}(2m+n)t + 3/4\,c\,V_m^2 V_n$$
$$\operatorname{Cos}(2m-n)t$$

Expanding the powers of Cosines (Cont'd.)

$+ 3/2\ c\ V_m V_n^2 + 3/4\ c\ V_m V_n^2 \cos(m+2n)t + 3/4\ c\ V_m V_n^2$

$\cos(m-2n)t$

$+ 3/2\ d\ V_m^3 V_n \cos(m+n)t + 3/2\ d\ V_m^3 V_n \cos(m-n)t$

$+ 1/2\ d\ V_m^3 V_n \cos(3m+n)t + 1/2\ d\ V_m^3 V_n \cos(3m-n)t$

$+ 3/2\ d\ V_m V_n^3 \cos(m+n)t + 3/2\ d\ V_m V_n^3 \cos(m-n)t$

$+ 1/2\ d\ V_m V_n^3 \cos(m+3n)t + 3/2\ d\ V_m V_n^3 \cos(m-3n)t$

$+ 15/4\ e\ V_m^3 V_n^2 \cos mt + 5/4\ e\ V_m^3 V_n^2 \cos 3mt$

$+ 15/8\ e\ V_m^3 V_n^2 \cos(m+2n)t + 15/8\ e\ V_m^3 V_n^2 \cos(m-2n)t$

$+ 5/8\ e\ V_m^3 V_n^2 \cos(3m+2n)t + 5/8\ e\ V_m^3 V_n^2 \cos(3m-2n)t$

$+ 15/4\ e\ V_m^2 V_n^3 \cos nt + 5/4\ e\ V_m^2 V_n^3 \cos 3nt$

$+ 15/8\ e\ V_m^2 V_n^3 \cos(2m+n)t + 15/8\ e\ V_m^2 V_n^3 \cos(2m-3n)t$

$+ 15/8\ e\ V_m^4 V_n \cos nt + 5/4\ e\ V_m^4 V_n \cos(2m+n)t$

$+ 5/4\ e\ V_m^4 V_n \cos(2m-n)t$

$+ 5/16\ e\ V_m^4 V_n \cos(4m+n)t + 5/16\ e\ V_m^4 V_n \cos(4m-n)t$

$+ 15/8\ e\ V_m V_n^4 \cos mt +$

$+ 5/4\ e\ V_m V_n^4 \cos(m+2n)t + 5/4\ e\ V_m V_n^4 \cos(m-2n)t$

$+ 5/16\ e\ V_m V_n^4 \cos(m+4n)t + 5/16\ e\ V_m V_n^4 \cos(m-4n)t$

Again Collecting Terms

$E_o = 1/2\ b\ V_m^2 + 1/2\ b V_n^2 + 3/8\ d\ V_m^4 + 3/8\ d\ V_n^4$

$+ 3/2\ c\ V_m^2 V_n + 3/2\ d\ V_m^2 V_n^2 + 3/2\ c\ V_m V_n^2$

$+ \cos mt\ (a\ V_m + 3/4\ c\ V_m^3 + 5/8\ e\ V_m^5 + 15/4\ e\ V_m^3 V_n^2$

$+ 15/8\ e\ V_m V_n^4)$

$+ \cos nt\ (a\ V_n + 3/4\ c\ V_n^3 + 5/8\ e\ V_n^5 + 15/4\ e\ V_m^2 V_n^3$

$+ 15/8\ e\ V_m^4 V_n)$

$+ \cos 2mt\ (1/2\ b\ V_m^2 + 1/2\ d\ V_m^4 + 3/2\ d\ V_m^2 V_n^2)$

Again Collecting Terms (Cont'd.)

$+ \text{Cos}\,2nt\,(1/2\ b\ V_n^2 + 1/2\ d\ V_n^4 + 3/2\ d\ V_m^2\ V_n^2)$

$+ \text{Cos}\,3mt\,(1/4\ c\ V_m^3 + 5/16\ e\ V_m^5 + 5/4\ e\ V_m^3\ V_n^2)$

$+ \text{Cos}\,3nt\,(1/4\ c\ V_n^3 + 5/16\ e\ V_n^5 + 5/4\ e\ V_m^2\ V_n^3)$

$+ \text{Cos}\,4mt\,(1/8\ d\ V_m^4)$

$+ \text{Cos}\,4nt\,(1/8\ d\ V_n^4)$

$+ \text{Cos}\,5mt\,(1/16\ e\ V_m^5)$

$+ \text{Cos}\,5nt\,(1/16\ e\ V_n^5)$

$+ \text{Cos}(m+n)t\,(b\ V_m\ V_n + 3/2\ d\ V_m^3\ V_n)$

$+ \text{Cos}(m-n)t\,(b\ V_m\ V_n + 3/2\ d\ V_m^3\ V_n)$

$+ \text{Cos}\,2(m+n)t\,(3/4\ d\ V_m^2\ V_n^2)$

$+ \text{Cos}\,2(m-n)t\,(3/4\ d\ V_m^2\ V_n^2)$

$+ \text{Cos}(2m+n)t\,(3/4\ c\ V_m^2\ V_n + 15/8\ e\ V_m^2\ V_n^3 + 5/4\ e\ V_m^4\ V_n)$

$+ \text{Cos}(2m-n)t\,(3/4\ c\ V_m^2\ V_n + 15/8\ e\ V_m^2\ V_n^3 + 5/4\ e\ V_m^4\ V_n)$

$+ \text{Cos}(m+2n)t\,(3/4\ c\ V_m\ V_n^2 + 15/8\ e\ V_m^3\ V_n^2 + 5/4\ e\ V_m\ V_n^4)$

$+ \text{Cos}(m-2n)t\,(3/4\ c\ V_m\ V_n^2 + 15/8\ e\ V_m^3\ V_n^2 + 5/4\ e\ V_m\ V_n^4)$

$+ \text{Cos}(3m+n)t\,(1/2\ d\ V_m^3\ V_n)$

$+ \text{Cos}(3m-n)t\,(1/2\ d\ V_m^3\ V_n)$

$+ \text{Cos}(3m+2n)t\,(5/8\ e\ V_m^3\ V_n^2)$

$+ \text{Cos}(3m-2n)t\,(5/8\ e\ V_m^3\ V_n^2)$

$+ \text{Cos}(2m+3n)t\,(5/8\ e\ V_m^2\ V_n^3)$

$+ \text{Cos}(2m-3n)t\,(5/8\ e\ V_m^2\ V_n^3)$

$+ \text{Cos}(4m+n)t\,(5/16\ e\ V_m^4\ V_n)$

$+ \text{Cos}(4m-n)t\,(5/16\ e\ V_m^4\ V_n)$

$+ \text{Cos}(m+4n)t\,(5/16\ e\ V_m\ V_n^4)$

$+ \text{Cos}(m-4n)t\,(5/16\ e\ V_m\ V_n^4)$

APPENDIX IV

A SUMMARY OF EXPERIMENTAL WORK

The writer performed his industrial Experience Tour at the Hewlett-Packard Co., at 275 Page Mill Road, Palo Alto, California, during the months of January and February 1956. The work was performed under the supervision of Dr. Peter D. Lacy, who is the head of Development of Microwave Devices for the company.

The object of the work was to develop methods of using a traveling wave amplifier as a bistable oscillator using internal feedback loop(s) so arranged that it may be keyed from one frequency to the other in response to a pulse of sinusoidal voltage of appropriate frequency and of suitable shape and amplitude.

The -hp- 491A Traveling Wave Amplifier was the basic vehicle for the tests. This instrument is briefly described in appendix II.

The basic scheme, suggested by Dr. Lacy, was to place a regenerative loop on the amplifier between the normal input and output couplers, so proportioned that it would support two stable modes of oscillation. The leakage from the output end of the loop would provide an input signal for the output section of the tube, which would act as a normal amplifier with reduced gain and provide a useful signal at the normal output of the system. That portion of the tube from the normal input to the front of the regenerative loop, termed the forward gain section, would also operate as an amplifier with reduced gain, and provide a method of injecting a trigger signal at saturation amplitude into the loop.

The TWA is a distributed constant four-terminal active network providing gain as a linear function (approximately) of helix length on a decibel basis. Hence the recirculation point of view is appropriate in considering the regenerative circuit. Inasmuch as the total loop delay forms an upper limit on the rate of mode switching it is desirable that the loop be only as long as is necessary to provide adequate gain to sustain stable oscillation. This is the basic reason for placing the loop internal to the amplifier circuit. The gain of this amplifier is approximately 10 db per inch of helix and the time delay is about two millimicroseconds per inch of helix. The latter corresponds to a helical propogation velocity of about 5% the speed of light based on the linear dimension.

The method of mode switching suggested by Dr. Lacy depends on the following mechanism: assume the two stable modes are f_1 and f_2 and that a keying signal $f_0 = f_1 + f_2$ is available. The system is oscillating at f_1 when f_0 is impressed on the loop by modulating the beam in the front gain section. Due to the non-linear action of the system at high level the difference frequency $f_0 - f_1 = f_2$ will be produced. With correctly chosen shape, length and amplitude of the signal at f_0 and an appropriate non-linearity it is possible to produce a magnitude coefficient for f_2 that is larger than that of f_1, and when the signal at f_0 is withdrawn the system will be excited more strongly at f_2 than at f_1. If the two modes are equally preferred switching will be accomplished.

A block diagram of the system described above is shown below:

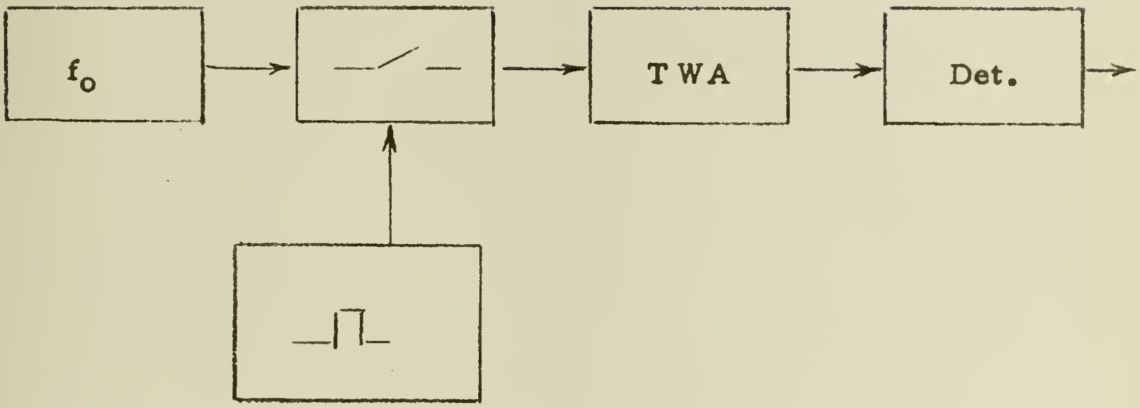


Figure 1. System Diagram

The signal at f_0 could be supplied by a small klystron, or from a signal generator followed by a TWA. A special microwave switch available at the laboratory would be necessary to produce an approximately rectangular pulse at f_0 of duration in the order of millimicroseconds. A special pulse-forming system would be necessary to key the switch. The detector could be any device capable of determining which of the two normal modes were operating, such as a microwave discriminator.

Evidently the logical first step in the procedure was to modify the traveling-wave amplifier to obtain bistable oscillation, and this course was pursued.

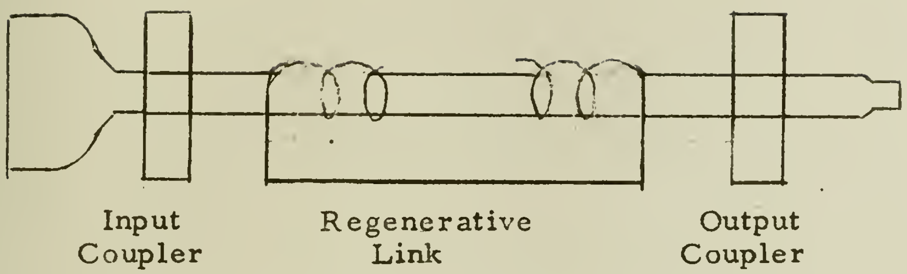


Figure 2. Regenerative Circuit Schematic

The first scheme attempted to give the desired regeneration was to employ a length of standard helix used in making the couplers, forming a link with the output coupled back to the input by a single wire. This is shown schematically in Figure 2. A modified -hp-809B Slotted Line suitable for inserting the traveling wave tube and exciting the helix was employed to measure the VSWR on the helix due to the presence of the coupler in an effort to determine its influence on the helix propagation. Difficulty was encountered because both a helical and a coaxial wave were propagated, and the standing wave pattern could not be interpreted. A magnetic probe was constructed in an effort to separate the field components and thereby resolve the standing wave into two waves. Since it was not feasible to provide a Faraday shield for the loop we succeeded only in complicating the problem by combining electrostatic and electromagnetic coupling. Finally an estimate of the number of turns needed in each helix was obtained and the loop constructed on the basis of this information and installed in the amplifier capsule.

The dimensions and arrangement of the assembly are shown below.

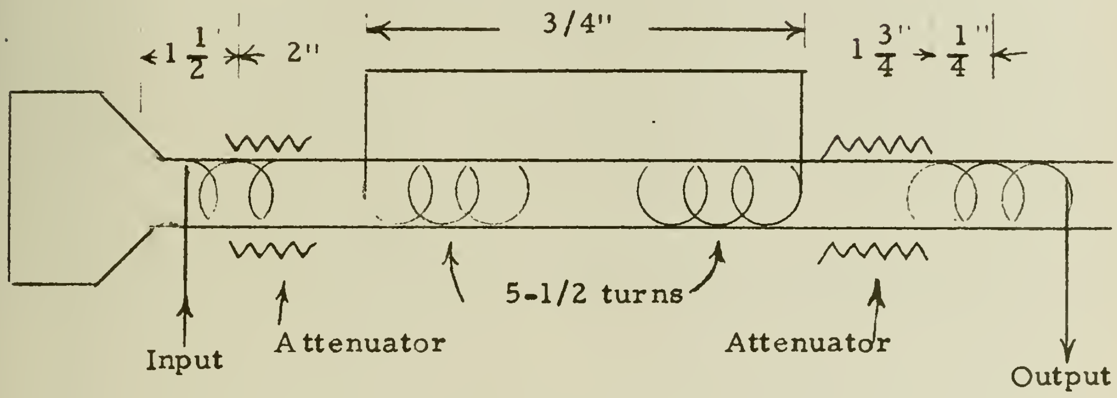


Figure 3. Schematic of Installed Circuit

No oscillation was sustained at maximum gain, and measurement and plotting of gain over the normal frequency range (2.0-4.0 KMC) showed no evidence of regeneration in an orderly manner, but rather had the influence of causing the gain curve to fluctuate erratically. One or two variations were tried and this method was abandoned in favor of using two standard couplers connected with a 50 ohm coaxial line.

The standard couplers are shielded helices with the shields in the form of a bobbin, the flanges forming supports to fit inside the capsule. The shielded helix has a nominal impedance of 50 ohms when coupled to the tube helix. The nether end of the helix is terminated in a 53 ohm resistor to prevent or reduce reflection of energy not coupled into the inner (tube) helix.

The two couplers were connected with a copper tube coax using teflon dielectric with a two-inch length (outside dimension) and installed in the capsule. Oscillation was obtained on three different frequencies with appropriate helix voltage, and the beam current (20 ma normal) was reduced to seven milliamperes to prevent oscillation and allow gain measurement. The gain was measured every 50 mcs from 1000-4000 mcs with helix voltages 750-1150 volts in 100 volt steps. The plots showed ample evidence of spurious regeneration that could not be accounted for by the presence of the intentional loop.

When the gain was increased to produce oscillation, the condition of oscillation was found to be sensitive to input and output impedance. A crystal rectifier at the input terminal showed backward wave power. It was further established that with a change in the helix voltage the output crystal current abruptly increased at the same time the input terminal crystal current went to zero. This corresponded to a shift in the output frequency from 1910 to 3220 mcs, or vice versa, depending on the direction in which the helix voltage was being changed. There was evidence of hysteresis, or frequency pulling, until the signal was pulled out of the mode.

These observations indicated that there were at least two regenerative loops influencing conditions, and possibly more. Inasmuch as the coupling between the beam and the tube helix is quite loose, it is not possible to remove modulation from the beam, so that any modulation influences conditions in all portions of the tube between the point of origin and the output. Because, though the beam passes through a coupler subsequent to modulation, all that can occur even with perfect coupling, is that the helix wave corresponding to this modulation would be removed, only to be re-excited by the modulation on the beam. Anything coupled to the tube helix offers an opportunity for impedance discontinuities with consequent fractional reflections of the helix wave at that point. This becomes a backward wave on the helix which will suffer little loss except cold-loss deliberately coupled into the helix. In this way the possibility of several regenerative loops becomes evident.

In view of the foregoing, some method of measurement was needed to allow determination of conditions inside the loop. We decided to construct and install on the coax feedback line a directional coupler with the ends of the coupler line accessible outside the capsule. A coupler with a nominal -20 db coupling was constructed as per the sketch in figure 4 below. It was duly installed with considerable mechanical difficulty, inasmuch as the line from the input coupler as well as the two additional ones from the directional coupler had to pass through the flanges of the output coupler

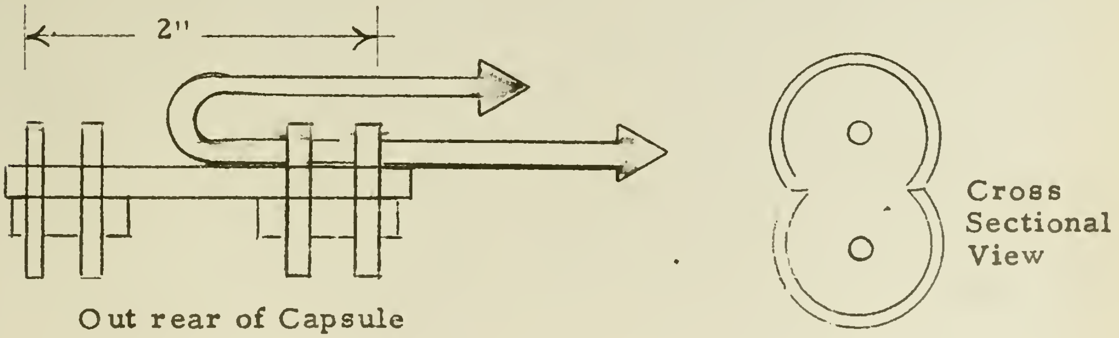


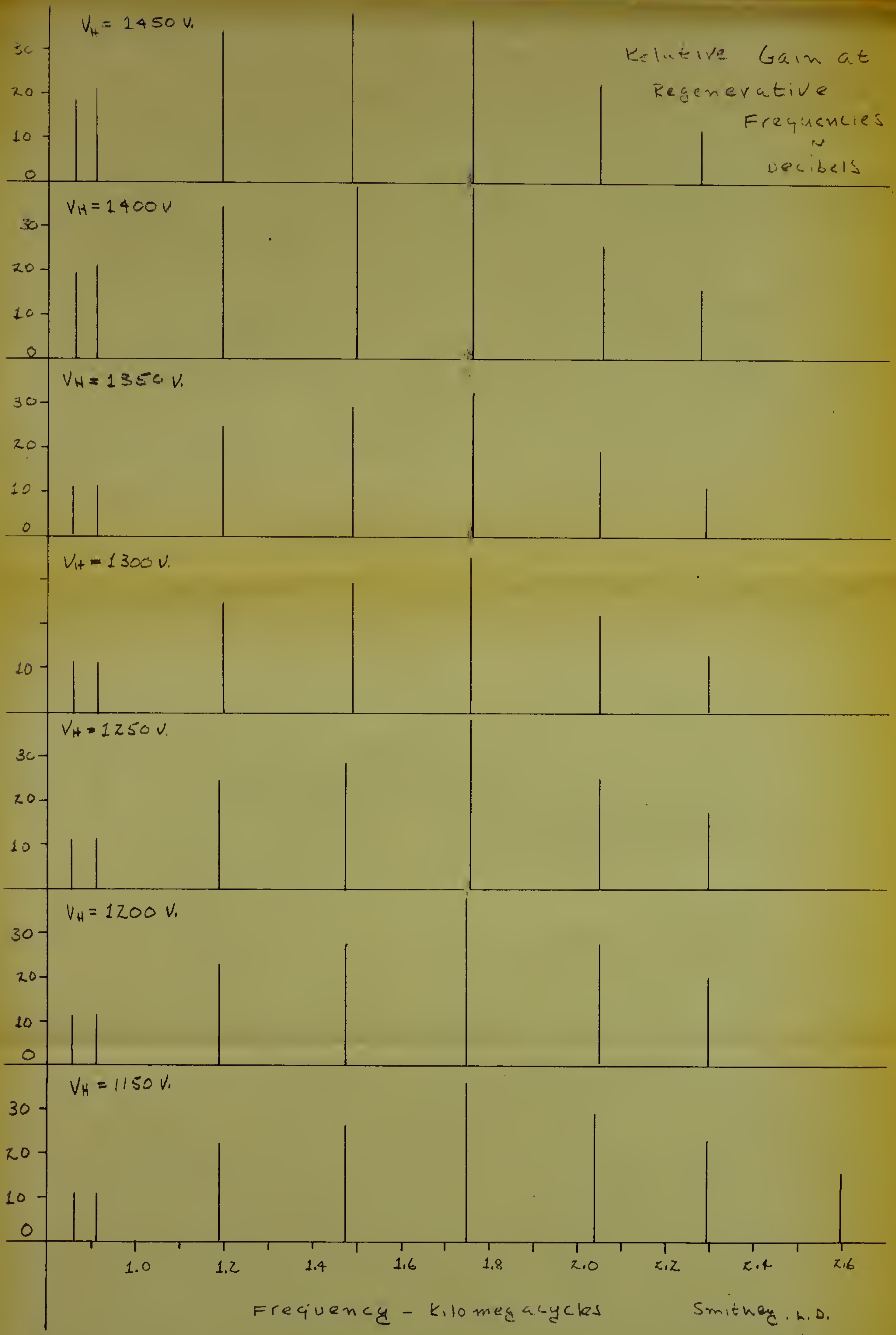
Figure 4. Directional Coupler used on Regenerative Loop

thence out the rear of the capsule. To make the situation more difficult, all these things had to be installed through a hole in the side of the capsule, a feat not possible without bending the lines in the process. It was then impossible to straighten them properly after they were inside the capsule. The results were rewarding, however.

In conjunction with this installation an addition was made to the anode supply regulator to permit square wave modulation of the beam current, and thus system gain, so that the HP-415A VSWR indicator could be used to measure what would otherwise be a CW signal. The details are shown in appendix III.

The coupler was found to work quite effectively inasmuch as the coupling was small enough that conditions in the loop were not materially affected, and signals available therefore were entirely shielded from all other parts of the circuit, so that a direct indication of what was occurring in the loop was available. In addition, it was now possible to inject a signal directly into the loop where heretofore it had been necessary to inject the signal through the front gain section of the TWA. We also determined the VSWR in the feedback loop to be at an acceptably low level.

Measurements made through the directional coupler allowed construction of the histrograms similar to those of Figure 5, with helix voltage as a parameter, and reasonable assurance that the regeneration was due to the loop under consideration. These data also allow the determination of the loop delay as approximately four millimicroseconds from the relation $\tau = 1/\delta f$. These results were quite


Relative Gain at Regenerative Frequencies in Decibels
V_H = 1450 V.
V_H = 1400 V
V_H = 1350 V.
V_H = 1300 V.
V_H = 1250 V.
V_H = 1200 V.
V_H = 1150 V.
30
20
10
0
1.0
1.2
1.4
1.6
1.8
2.0
2.2
2.4
2.6
Frequency - kilomegacycles
Smithey, L.D.
1-25-56
Figure

gratifying if not of monumental importance, since it was the first instance in which the quantity of interest could be separated from extraneous ones. It should be noted that these regenerative gain magnitudes have an envelope corresponding approximately to the normal amplifier gain curve, as would be expected.

It can be deduced also from these data that the frequency separation is not as great as is desired for bistable operation, but greater separation requires a smaller τ, which means a shorter loop and consequently less loop gain. This in turn means relatively greater significance attached to spurious regeneration and less beam saturation due to the sustained oscillation. The former may easily preclude operation entirely, while the latter may not permit switching or at best will contribute to mode instability.

On the first attempt after installing the coupler there was evidence of instability in the front gain section, and a histogram showed spurious responses. The capsule was opened up and the tube removed and a longer attenuator installed on the front gain section. This stabilized the system and the spurious gain peaks disappeared. Higher helix voltages than normal were used (above 1050V.) to investigate the response in the phase dispersive region of the frequency spectrum. It is noticed from the histogram that the amplification band becomes narrower and peaks at a lower frequency. This feature offers some discrimination to adjacent regenerative frequencies and eases the problem of mode selection. Not sufficiently, however.

It was next considered desirable to shorten the loop and consequent time delay to increase the value of δf. To do this the directional coupler had to be removed, and was not re-installed in view of the mechanical difficulty involved. Two coupler shields had one flange each removed and the bushing milled on the end from which the flange was removed to shorten the shield by approximately one eighth inch to provide the maximum of exposed tube helix for a given overall loop dimension. This was considered feasible since the shield is somewhat longer than the helix inside. The two couplers were than connected together with a piece of coax line so that the overall length was 1-1/2 inches, leaving 5/8 inch of exposed tube helix between them. The assembly was then installed in the capsule using a 2-1/8 inch helical attenuator forward and a 1-1/4 inch attenuator on the output section composed of half aquadag and half helical attenuator.

The results obtained from this loop were quite unsatisfactory. There was evidence of oscillation associated with the front section, and the loop gain had been reduced to the point that it was necessary to increase the beam current from about four to about 12 milliamperes. By moving the loop assembly forward to reduce the length of the front section and increasing the attenuation on the output section,the system was stabilized considerably but not adequately. The only gratifying result was that the frequency spacing was increased from about 260 mcs to 380 mcs. The conclusion reached, among others, is that impedance relations were seriously disturbed by reducing the length of the shields.

When the capsule was opened up the terminating resistor (53.3 ohms, 1/8 watt precision) for the forward coupler of the regenerative loop, which forms the terminating load for the front gain section, was found to be charred from over-heating. This indicated that the forward gain section had been oscillating at a quite high level.

The regenerative couplers were replaced with standard ones without alteration, using the same overall dimension as that for the modified ones. Two inch attenuators composed of half aquadag and half helical material were installed forward and aft. When the beam current was increased to the point of oscillation power was available at the input and output, both at the same frequency. A large number of apparently unrelated regenerations were observed at low gain. The gain was observed to be about equal at 1010 mcs and 1670 mcs with helix at 1170 volts, but 1670 was the preferred mode, for each time the system was excited at low gain on 1010 mcs and then the gain increased it would switch abruptly to 1670 mcs. There was some evidence that the mode strength (relative) was dependent on beam current. This had been noticed on several previous occasions.

Under the conditions described above with 26 milliamperes and 1100 volts on the helix, twenty-six frequencies were measured in the output at various levels, and indications of several others were observed that were too weak to measure. This was obviously an untenable situation.

Although not previously mentioned, there had been indications all along that a higher degree of beam saturation was needed than was being achieved, in order to obtain the requisite non-linearity for switching. But the degree of control over reflections that was needed to obtain the necessary gain and stability together was not forthcoming.

As a means of increasing the regenerative frequency separation the use of coaxial band pass and/or band reject filters had been considered and hastily rejected in view of the space available. This being a problem still to be solved the notion that multiple feedback loops might be useful was entertained. After a little investigation two or three different configurations showed promise, and one was tried.

The multiple loop system deemed most feasible for a first attempt was chosen: first, the idealization of it shows some desirable properties and second, it may readily be realized in this system, at least to a first approximation, using only two gain sections, three couplers and two feedback lines.

How this configuration was realized is most easily shown with the aid of the diagram in Figure 7. The coupler k_2 employed a reduced number of turns so that power transfer was incomplete so that some of the power fed back from k_3 passed through k_2 to be received by k_1 and transferred to the tube helix. Also, only a fraction of the power produced by μ_1 was collected by k_2 to be fed back to k_1, while the remainder became an input for μ_2. It can easily be shown that there is a discrete set of frequencies at which all three loops will be regenerative simultaneously and that a set of such frequencies will be preferred oscillatory modes.

The regenerative loops were constructed using standard couplers with the number of helical turns reduced appropriately on the center one. Inasmuch as the desired properties require $\beta_1 = \beta_2$, $\mu_1 = \mu_2$ the two gain sections were made of different lengths and the longer one appropriately attenuated.

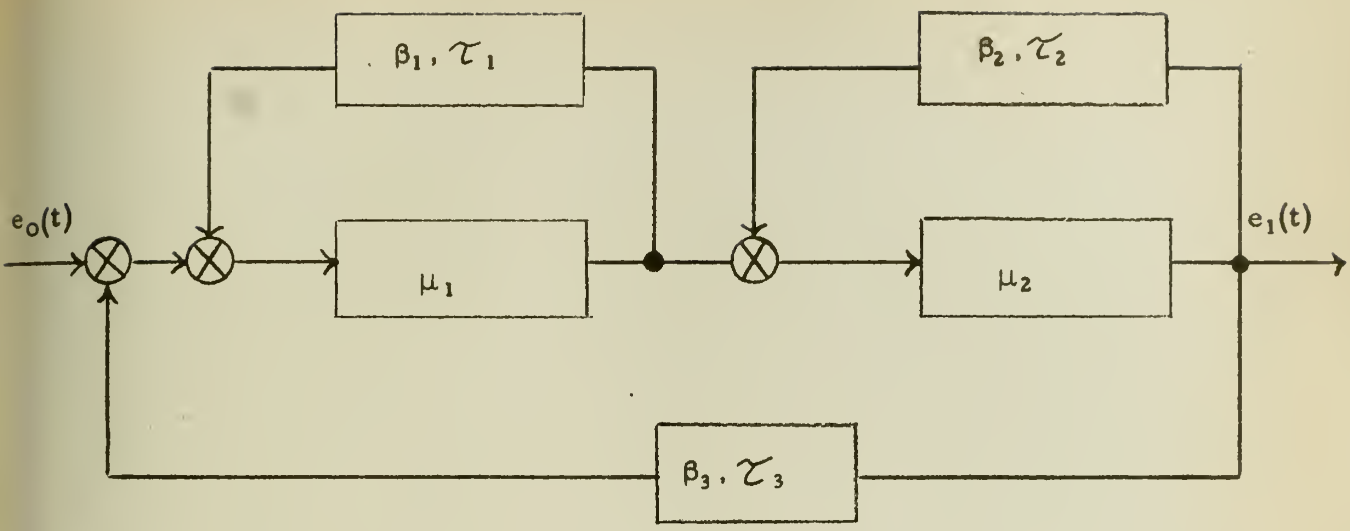


Figure 6. Multiple Loop Diagram

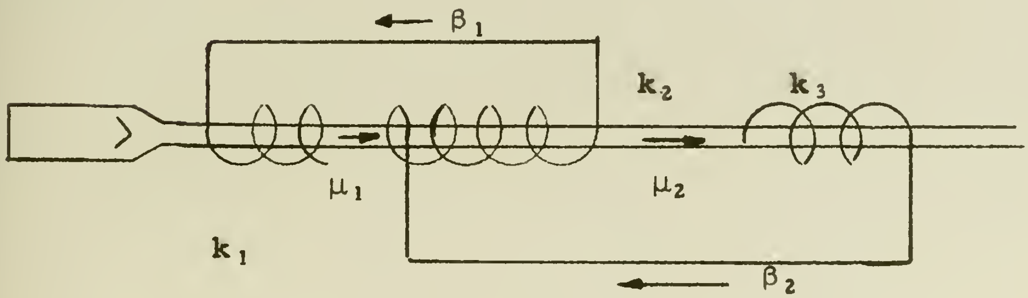


Figure 7. Circuit Realization

The systems oscillated, to be sure, but showed more spurious regenerations than any system tried so far. In particular, it showed frequencies regularly spaced at 40 mcs, corresponding to a loop delay of 25 mμsec., about twice the slow wave time once down the tube, indicating that the total helix between the input and output couplers was included in one regenerative loop with slow wave propagation each way. This may be attributed, at least in part, to the fact that insufficient attenuation could be placed on the helix. With each gain section progressively excited by beam modulation that cannot be removed by the couplers, there must have been something upward of 30 db gain that was effective for a forward wave excited in the lead gain section. The lack of attenuation on the helix allowed energy reflected from the output to travel back up the helix to the input, thus closing the loop. The effect of deliberate regeneration was effectively submerged.

Being now more or less at loss for something constructive to do, the thought of determining the results of deliberate reflections was entertained. A single coupler was equipped with a five turn helix terminated in a copper sleeve; the assembly was inserted in the coupler sleeve, with the helix forward. The coupler was then placed in the capsule up against the fast wave suppressor, leaving about 1-3/4" of helix exposed between this and the input coupler. The reflection was nearly complete, and the system oscillated, but of course the termination of the input line (the -hp- 614 or 616 generator in this case) was exposed to considerable power flowing in the reverse direction unless this too was deliberately reflected. The frequency and amplitude of the oscillation were sensitive to input termination and to beam current as well as to helix voltage.

A review of all previous results convinced us that our difficulties could be traced, for the most part, to spurious feedback involving the front gain section. It seemed necessary to stabilize the tube against this type of regeneration, and thus reducing the front section gain and/or the back wave attenuation in this section. Accomplishment of this end seemed to require as long a forward gain section as possible to allow a maximum of attenuation. This seemed reasonable, for the gain over the band of interest could be limited to a few decibels by the use of attenuators, and the longer this attenuator the greater would be the attenuation of the backward wave. This suggested that the normal output coupler be one of those in the feedback loop.

A new (incomplete) capsule with no stop rings was obtained to allow the couplers to be inserted from the rear end. A stop was installed for the front coupler. A pair of new couplers were made up, and installed. The tube was then inserted, using a production attenuator, to test the couplers. The forward coupler differed from standard only in that both ends of the helix were terminated in a 50 ohm coax; both lines were brought out the rear of the capsule.

The tube was inserted and the input line was checked with the -hp- 416 reflectometer with a cold helix when the other coax was terminated. It was found that the reflection averaged 30% when a 5" aquadag attenuator was used on the tube. This was not as flat as desired but deemed satisfactory for the impending trials.

Next the capsule was assembled as an amplifier using a 6" aquadag attenuator and the gain curve of Figure 8 was obtained. The

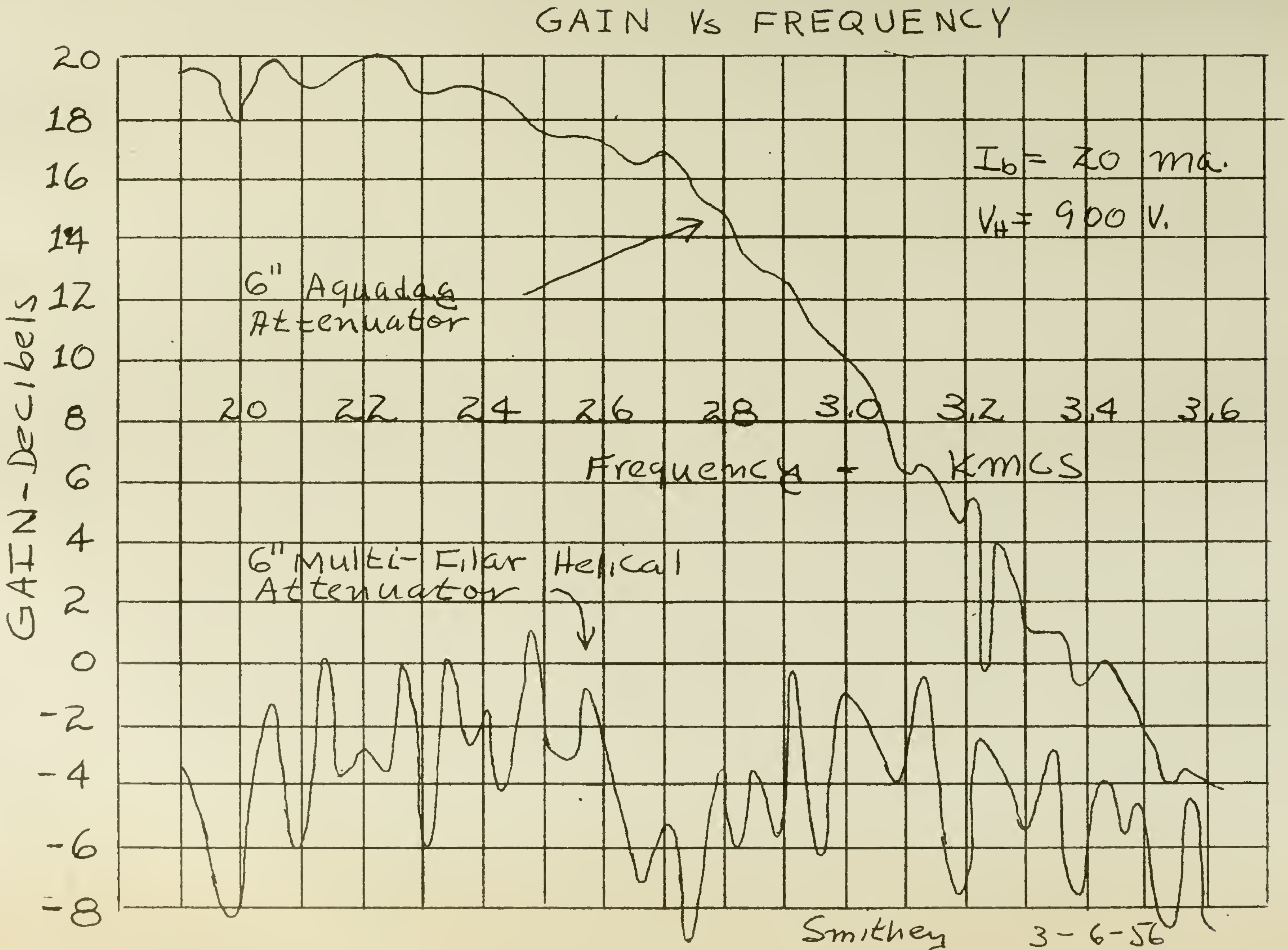

GAIN Vs FREQUENCY
GAIN-Decibels
20
18
16
14
12
10
8
6
4
2
0
-2
-4
-6
-8
2.0
2.2
2.4
2.6
2.8
3.0
3.2
3.4
3.6
Frequency - KMCS
Ib = 20 ma.
VH = 900 V.
6" Aquadag Attenuator
6" Multi-Filar Helical Attenuator
Smithey 3-6-56

Figure 8

attenuator was replaced with a 6" helical attenuator and another curve was run for comparison. It is also shown in Figure 8.

The ripple in the gain curve using a helical attenuator was believed caused, at least in part, by the impedance discontinuity caused by the end of the attenuator. For this reason it was decided to use a short length of aquadag attenuator in conjunction therewith in the final assembly. This short section was tapered to four points in a half inch to further reduce these reflections.

Next the regenerative loop with two inch outside dimension was made up, using the output coupler. The capsule was then assembled as shown in Figure 9. It is to be noted that the two free ends of the helices in the loop couplers were terminated in coax lines and brought out the rear for external loading and/or measurement. The dimension from the rear side of the output coupler to the end of the tube helix was 5/16".

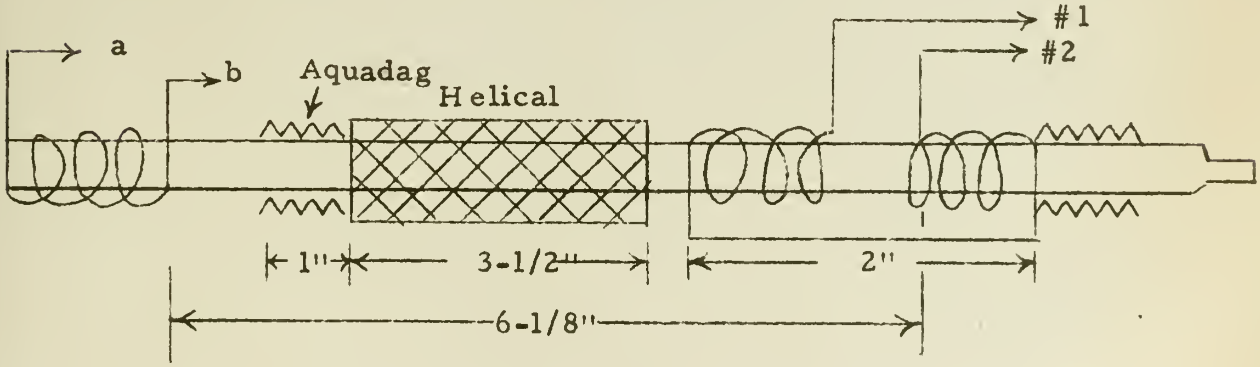


Figure 9. Assembly Dimensions

The system oscillated gingerly, and crystal rectifiers failed to show power on lines a and b (to the input coupler), but we were surprised to find crystal current from line #2 to be 18 ma compared with 4.5 ma on line #1; this indicated a strong backward wave was coupled into the rear helix from somewhere. When the aquadag attenuator was removed from the tail end of the tube current #2 went down to 6.5 ma, while that of #1 went up to 5.5 ma. A 1" helical attenuator caused the same set of conditions as did the aquadag one. When the aquadag one was tapered to three points in a half inch the conditions were $I_1 = 5.8$ ma and $I_2 = 8.8$ ma. Under the latter condition the oscillation was at 2100 mcs, but frequencies were indicated each 120 mcs from 1600 to 2220 mcs.

The $\delta f = 120$ mcs indicated $\tau = 8$ mμsec, about twice that accounted for by a 2" loop with a coax feedback, indicating the loop was closed by slow wave reflections, possibly from points outside the couplers. It was found that when the coupler assembly was moved toward the end of the tube until the rear flange of the coupler was coincident with the end of the tube helix that $I_1 = 5.7$ ma, $I_2 = 13$ ma and $\tau = 6.7$ mμsec.

The plot of relative regenerative gain magnitudes corresponding to the above conditions for one helix voltage is shown in Figure 10. The normal input was used for obtaining this curve, and it was very difficult to get enough signal into it, even when using an -hp- 491A amplifier, due to low gain (or perhaps loss) in the front section. We elected to again determine these magnitudes using the front end of the rear coupler as an input, and using the rear end of the center coupler as an output.

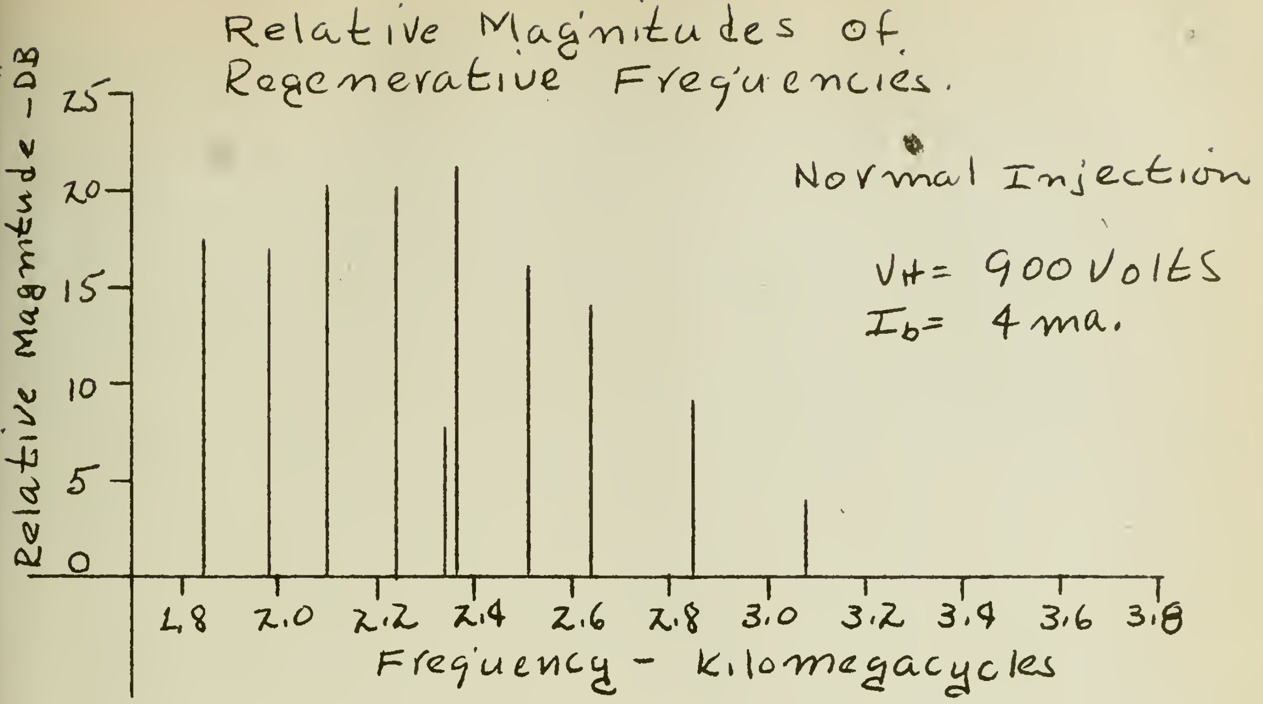


Figure 10

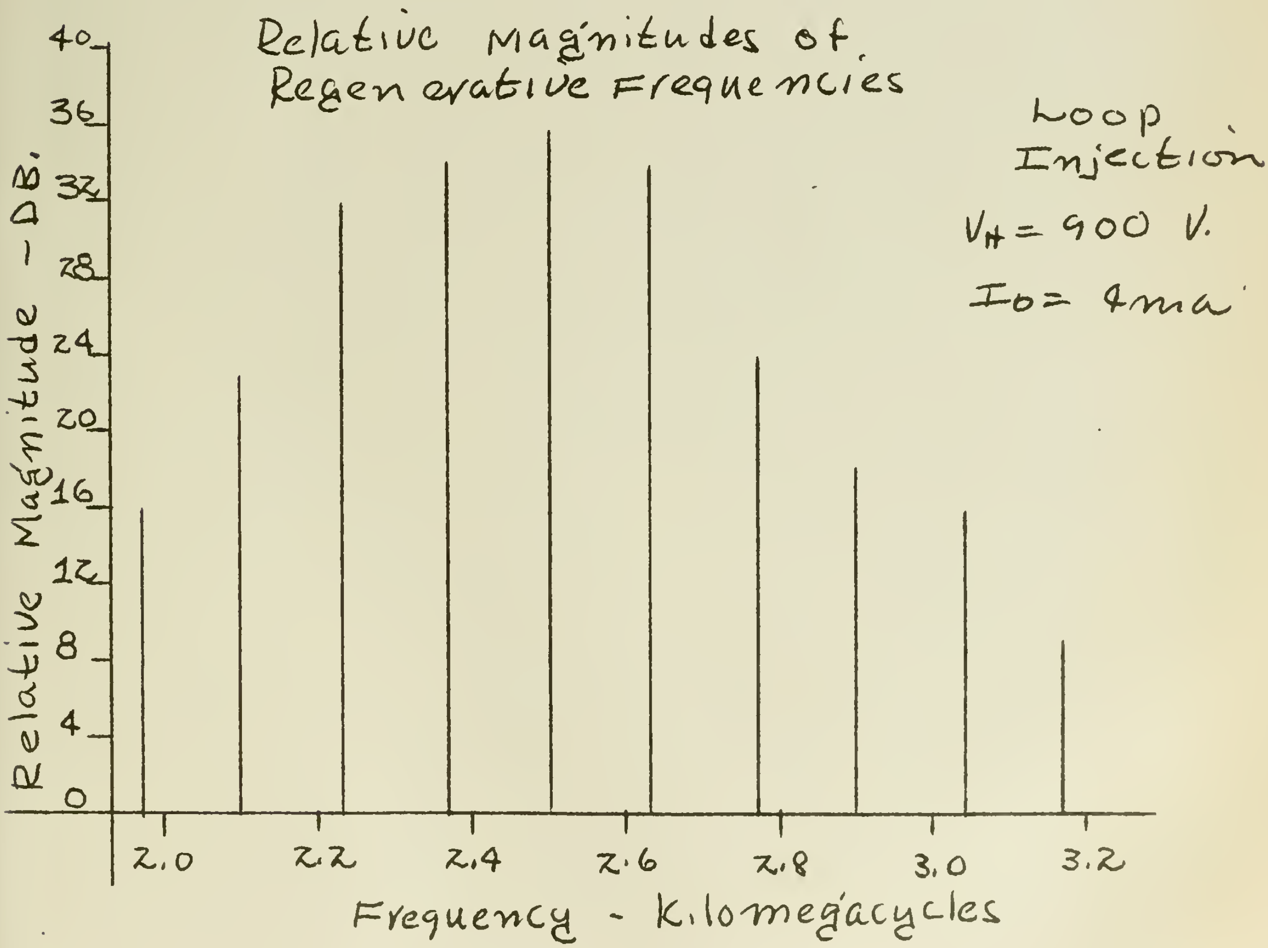


Figure 11

This plot is shown in Figure 11 for comparison with that of Figure 10.

Various standing wave patterns were made, without yielding any significant information, except that there was a considerable power leakage past the couplers, enhancing reflections; the leakage was greater for higher degrees of beam saturation, apparently.

It is significant that the effect of the deliberate loop could only be detected indirectly, and that the unintentional one largely determined the circuit characteristics.

This configuration was examined for conditions of bi-stability using a single shot pulser to modulate the beam current. We were pleasantly surprised to learn that the system was bi-stable, as indicated by the table below:

V_h	825	910	980	1110	1270	volts
I_b	10	16	16	18	18	ma
f	2590	2590	2080	1840	1700	mcs
	2710	2710	2200	2080	1840	

These combinations were investigated in various ways. We found that when the beam current was modulated at 5000 cps with 10 μsec pulses that conditions could be adjusted so that the modes were equally probable (approximately). Mode detection could be easily accomplished by use of the rectified output and a dc scope, since the magnitudes of the modes differed slightly, giving distinct traces on the screen. That each was excited approximately half the time was indicated by equal trace brightness. That the system was stable in mode was shown by using single shot modulation and noting that the output remained at constant level. Also, single trace photos were taken using 50 cycle modulation showing that the mode was maintained until the oscillator was disabled. There was no evidence of the type of instabilities reported by the Stanford group.

The best pair of modes proved to be 2590-2710 mcs with V_h = 910 volts and I_b = 16 milliamperes, from the standpoint of being less critical of adjustment. Conditions were obtained supporting simultaneously three, four and five modes. The latter was very critical of adjustment. These were not pursued further, since bi-stability was of interest.

The front plate of the amplifier was removed, and a 47 ohm resistor was installed in series with the helix lead at the tube socket. A coax line was attached thereto, with the center conductor connected to the helix side. This allowed application of a pulse in series with the helix to permit phase modulation.

The transmission line pulser was used to apply pulses to the helix in lengths varying from five to 100 mμsec and of magnitudes from zero to 250 volts peak. Effects of the longer pulses could be detected as a change in the output amplitude, but no tangible influence was noted for pulses of duration less than about 20 mμsec. No switching was accomplished in this way. This effort is not considered conclusive, for there is some doubt that the short pulses actually got on the helix in view of the method or wiring, and that for this purpose the helix is essentially an open-ended single-wire transmission line. The total electrical length of the helix is about 12 mμsec; reflections surely occured, and the pulses in this order of length may have been lost in multiple reflections.

A balanced microwave modulator was available for use as a microwave switch. This device has the property that it may be balanced for zero output with CW input, when its two crystal rectifiers have zero bias. The crystals are oppositely polarized so that when a pulse is applied to them in parallel one is made to conduct heavily and the other is cut off. This unbalanced the system and an output is obtained for the duration of the pulse. CW pulses are obtainable of precisely defined interval with envelope rise time of about one mμsec. Unfortunately, this device may only be balanced over the range from 2200 - 3200 mcs.

An attempt was made to pulse the input of the system to obtain switching. The modes 2590 and 2710 mcs were used, and the pulse was varied in the vicinity of the mean of these. The pulse out of the modulator was amplified with an -hp- 491A amplifier so that pulses up to about one watt level were available. Pulse lengths from five to twelve mμsec were tried at varying levels and frequencies. Erratic switching was obtained at the highest pulse level available and pulse duration of about 10 - 12 mμsec. Consistent switching was not obtained.

An array of equipment was then employed to obtain a pulse of frequency $f_0 = f_1 + f_2$ so that frequency doubling on the beam need not be relied upon. The pulse at the mean frequency from the modulator was amplified with an -hp- 490A amplifier and applied to a G-band waveguide through a coax line having the center conductor connected to the outer conductor through a microwave crystal for harmonic generation.

The waveguide cut-off frequency is about 3200 mcs, so the fundamental component was removed. The second harmonic was retained and higher harmonics were above the amplifier pass band. The pulse was subsequently amplified by a C-band amplifier and an -hp- 491A amplifier so that a pulse level of about one watt was available. The pulse fidelity was retained, but an inordinate amount of noise was obtained in the process. Oscillator stability was destroyed when this signal was applied to the input.

It was established, however, that when an rf pulse at a mode frequency was injected that the system was invariably switched from the opposite mode. This proves at least that switching by beam modulation is possible, if there were ever any doubt of it. The photographs made of this phenomenon also demonstrate that oscillation is being sustained at below maximum power beam saturation, for the power was shown to increase during the pulse (1 to 10 μsec duration and $I_b = 13$ ma). It was possible to increase the input pulse to and beyond that corresponding to maximum power output, for the power during the pulse could be made less than mode power. Switching was reliably accomplished under these conditions.