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SIMULATION METHODS APPLIED TO SINGLE-TUBE HARMONIC GENERATOR DESIGN

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Abstract iv

Problem Status iv

Authorization iv

INTRODUCTION 1

THEORY OF SIMULATION 2

SIMILITUDE vs ANALYTICAL APPROACH 4

LIMITATIONS IMPOSED BY INPUT CAPACITY OF THE STAGE 8

FEEDBACK CONSIDERATIONS 9

COMMENTS ON MODULATION 10

CONCLUSIONS AND REMARKS 10

RECOMMENDATIONS 11

ABSTRACT

A method has been found for improving the gain obtained in frequency multipliers through a study of the process of harmonic generation in single-tube circuits. Since distorted waveforms having a fundamental above 500 kc, and prominent harmonic components up to the 10th, could not be observed with fidelity, simulation methods were used. Circuits operating in the audio range were used to study distorted waveforms having the desired shapes. In this way waveforms were produced in the audio-frequency circuits having shapes exactly like those to be expected in frequency multipliers designed to operate at much higher frequencies. All components were properly scaled in electrical size.

The conditions to be imposed upon the circuits in order to secure adequate simulation are derived. Experimental data is given for a specific simulation problem. Oscillograms showing the distorted waveforms under various conditions are presented.

The emphasis in this work has been directed at securing unity voltage gain utilizing a single pentode tube to obtain a frequency multiplication of 10.

PROBLEM STATUS

The progress reported here is a small part of the Electronic Standards Problem No. A-262. No further theoretical and similitude work will be done on frequency multipliers. A report on the actual full-scale multiplier construction will be given later. Work on the remainder of Problem No. A-262 will continue.

AUTHORIZATION

NRL Problem No. R10-23D (BuAer A-262)

SIMULATION METHODS APPLIED TO SINGLE-TUBE HARMONIC GENERATOR DESIGN

INTRODUCTION

Problem R10-23 (A-262R-U) requires frequency measurement facilities of high precision up to the microwave region, at the moment through the X-band. Contemplated standard-frequency sources which will operate at one, ten, one-hundred, and one-thousand Mc will require the following frequency multipliers:

(a) 100 kc to 1 Mc

(b) 1 Mc to 10 Mc

(c). 10 Mc to 100 Mc, and

(d) 100 Mc to 1000 Mc

The above tabulation of multiplier requirements makes those capable of multiplying by ten of particular interest, especially those single-tube designs where the output-input voltage ratio is unity.

Upon considering the frequency range of required multiplier operation, it was felt that conventional clipping methods employing resistor-capacitor components were not applicable over the entire range. Consequently, it was decided that resonant distorter circuits employing L, C, and R components would be used to provide suitable transients in the control-grid circuit of a multiplying tube.

An earlier report[1] shows that simple harmonic generators could be constructed which would supply an output-input voltage ratio of unity, provided a gain of 100 was obtainable at the frequency in question. In principle, the harmonic content of the grid input signal is produced by distorting a sine-wave driving signal by utilizing the sharp cut-off properties of a pentode, and by controlling grid current flow. The first efforts using this principle resulted in a satisfactory multiplier for converting 100 kc to 1 Mc where all waveshapes were readily observable on an oscilloscope having 2-Mc band-pass characteristics.

Considering the required range of frequency multiplier operation set forth previously, ordinary oscilloscope limitations would make observations of distorted wave forms of 1-Mc and 10-Mc fundamentals very unsatisfactory. Oscilloscopes capable of giving a true presentation of a distorted 1-Mc signal possessing prominent harmonics up to 10 Mc are

[1] H. H. Grimm, Single tube harmonic generator design, NRL Report No. R-3166, Sept. 1947

not common. Coupling to the circuits in which the distorted signal exists presents additional difficulties. Fast sweeps of the proper order of magnitude, 2 or 3 in./μsec, are obtainable, but suitable synchronization is difficult. Since the design of oscilloscopes suitable for this application is a formidable problem a method was sought for avoiding this difficulty.

Experimental attack by similitude methods offers an attractive solution because the higher-frequency multipliers could be scaled down to a unit, multiplying, for example, 1 kc to 10 kc. With reference to the four required multiplying ranges (a), (b), (c), and (d) previously set forth, it should be emphasized that range (a) requires no similitude attack; ranges (b) and (c) can be simulated, and range (d) has not been simulated. Plans call for the use of a circuit using lighthouse-type triodes for this purpose.

This report deals with the similitude principle as applied to the frequency ranges (b) and (c). In particular, the example simulation detailed in this report applies specifically to the frequency multiplier range (b).

THEORY OF SIMULATION

The simulation theory involved is very simple. The sine-wave input is to be seriously distorted by use of a non-linear impedance in the grid circuit of a tube. The important non-linearity is the sharp change in input admittance of the tube when grid conduction starts. Under conditions where transit time and feedback may be ignored, the non-linearity is in the resistive part of the tube admittance.

Tube design dictates the available tube input admittance values. The first requirement for adequate simulation at a convenient low frequency is the selection of vacuum tubes with small transit time and feedback effects at the output frequencies given by the multiplier ranges (a), (b), (c), and (d).

The conditions for correct simulation can be secured directly from the differential equations of the system. The equivalent distorter circuit is shown in Figure 1.

The subscript o denotes components in the multipliers a, b, c, and d. The subscript 1 denotes the corresponding quantities in the simulation circuit.

The differential equation for the circuit, Figure 1, is

$$C\,\frac{de}{dt} + Ge + \frac{1}{L}\int e\,dt = I\,\epsilon^{\,j\omega t}$$

and from this the steady-state solutions (subscript s), corresponding to the two circuits, actual and simulated, are:

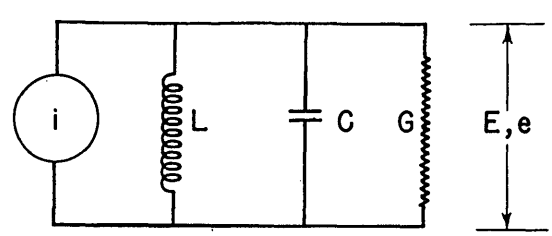

Fig. 1 - Equivalent Circuit for Simulation Theory

$$E_{so} = \frac{I_{so}}{j\omega_o C_o + G_o + \frac{1}{j\omega_o L_o}}$$

$$E_{s1} = \frac{I_{s1}}{j\omega_1 C_1 + G_1 + \frac{1}{j\omega_1 L_1}}$$

Now these steady-state solutions are identical if

$$I_{s1} = I_{so} \text{ ---(1)}$$

$$G_1 = G_o \text{ -----(2)}$$

$$\omega_1 C_1 = \omega_o C_o \text{ ---(3)}$$

$$\omega_1 L_1 = \omega_o L_o \text{ ---(4)}$$

Conditional equations of this sort establish the requirements which must be met in the similitude experiment. The first condition is satisfied by matching the amplitude and phase of the driving current. The second is dictated by tube design. The third and fourth give the frequency sensitive conditions that must be applied to the susceptance in the circuit. These relations can be summarized as:

$$\frac{\omega_1}{\omega_o} = \frac{f_1}{f_o} = \frac{C_o}{C_1} = \frac{L_o}{L_1}$$

The transient solutions (subscript t) are given by the equation

$$e = E_t \epsilon^{pt}$$

where

$$p = \frac{-G \pm \sqrt{G^2 - 4\frac{C}{L}}}{2C}$$

The two solutions, actual and simulated, are:

$$e_o = E_{ot} \epsilon^{\left(-\frac{G_o}{2C_o} \pm \frac{\sqrt{G_o^2 - 4\frac{C_o}{L_o}}}{2C_o}\right)t_o}$$

$$e_1 = E_{1t} \epsilon^{\left(-\frac{G_1}{2C_1} \pm \frac{\sqrt{G_1^2 - 4\frac{C_1}{L_1}}}{2C_1}\right)t_1}$$

These two solutions are identical ($e_o \equiv e_1$) if

$$E_{ot} = E_{1t} \text{ ----(5)}$$

$$\frac{G_o}{2C_o} t_o = \frac{G_1}{2C_1} t_1$$

and
$$\frac{\sqrt{G_o^2 - 4\frac{C_o}{L_o}}}{2C_o}\, t_o = \frac{\sqrt{G_1^2 - 4\frac{C_1}{L_1}}}{2C_1}\, t_1$$

The last equation can be put in the form

$$\left(\frac{t_o}{t_1}\right)^2 = \left(\frac{C_o}{C_1}\right)^2 \frac{\frac{G_1^2}{4} - \frac{C_1}{L_1}}{\frac{G_o^2}{4} - \frac{C_o}{L_o}}$$

This equality holds if

$$G_o = G_1 \text{ ------- } (6)$$

and

$$\frac{C_o}{C_1} = \frac{L_o}{L_1} = \frac{t_o}{t_1} \text{ --- } (7)$$

The ratio t_o/t_1 in the transient case is analagous to ω_1/ω_o in the steady stage case. Therefore the same conditions apply to both steady-state and transient cases. The non-dissipative components must be in the inverse ratio of the frequencies in the actual and simulated circuits. The component values given in Table I are experimental values employed to obtain performances in Figures 5 - 11. It is to be noted that these values comply with the foregoing theory. The value of ωL or ωC in one column is approximately equal to its partner in the other column. For the column headed 1.31 kc to 13.1 kc the capacities and inductances are 760 times those in the 1.0 to 10 Mc column. The frequency ratio $1.31 \text{ kc}/1.00 \cdot 10^3$ kc is 1/760.

SIMILITUDE vs ANALYTICAL APPROACH

Analytical treatment of different circuit arrangements can be obtained by using the methods developed by Guillemin and Rumsey for non-linear magnetic circuits.[2,3] An analysis of this type was employed in an exploratory fashion in the attack on this problem, but a treatment adequate for the subject purpose would require differential equations of one higher order than Guillemin and Rumsey used, and more numerous circuit components. Some analytical attempts, following Guillemin and Rumsey's method, were helpful in highlighting the qualitative aspects of this sort of transient problem, but it is believed that the similitude method gave experimental results much more rapidly than corresponding analytical answers could be obtained by known analytical methods.

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[2] E. A. Guillemin and P. T. Rumsey, Proc. IRE, Vol. 17, No. 4, pp. 629-651, April 1929.

[3] "Zur Theorie der Frequencvervielfachung durch Eisenkoppelung" Archiv für Electrotechnik, Band XVII; 1926.Heft 1 Sil 7.

TABLE I

Ratio Between the Component Sizes in the Simulation Circuit and the Actual Circuit

Ratio Between Component Sizes in the Two Columns, 760 to 1 Frequency Ratio 1 to 760

	1,310 kc to 13,100 kc	1.0 Mc to 10.0 Mc
L_1	13,500 μh	17.7 μh
C_1	1,100,000 $\mu\mu$f	1450 $\mu\mu$f
L_2	13,500 μh	17.7 μh
C_2	12,000 $\mu\mu$f	16 $\mu\mu$f
L_3	1,350 μh	1.77 μh
C_3	110,000 $\mu\mu$f	144 $\mu\mu$f
C_{pg}	20 $\mu\mu$f	.026 $\mu\mu$f

As shown in footnote (1), the distortion process, assuming trapezoidal distortion, results in an amplitude-loss factor, $C_n/E = 1/n^2$ where C_n is the amplitude of the nth harmonic component of a trapezoid pulse of amplitude A secured from a driving voltage E. Figure 2 gives a graphical presentation of these symbols.

The relationship $C_n = 0.5\,A/n$ was developed in footnote (1).

The voltage gain (for an output frequency 10 times the input frequency) required to maintain the input voltage level is 100. This large gain is feasible at 1-Mc output and also at 10-Mc output. At 100-Mc output, feedback and tube admittance conditions make it difficult to obtain a stable gain of 100. However, two stages of gain 10 are probably more practical. Nevertheless, it is possible that a neutralized stage with a gain of 100 could be built.

Miniature tubes (6AG5) with g_m values of 5000 microhms were used for simulation purposes. To obtain a gain (G) of 100, the antiresonant impedance (L_3C_3) Figure 3, of the plate circuit must be $Z_3 = G/gm = 100/5000 \cdot 10^{6} = 20{,}000$ ohms.

The capacity and inductance values in Table I were calculated, assuming a Q of 100 and an antiresonant impedance of approximately 20,000 ohms, from $X_3 = Q\omega L_3 = Q/\omega C_3$.

The coils of the transformers (GR type 107) used for simulation work did not have a Q = 100. The actual gain, at the output frequency was somewhere in the range 30 - 50. This decrease in gain, however, does not change the qualitative results obtained.

In the simulation experiments 6AG5 tubes were used, but for the actual multipliers, (b) and (c), 6AU6 tubes should be used because they have much lower Cpg. Both types have close spacing and small transit time under the subject experimental conditions.

A block diagram of the similitude experimental arrangement is shown in Figure 3, and the corresponding similitude circuit by that of Figure 4. The use of an electronic switch with the oscilloscope greatly facilitated the experimental work because the presentation of two waveforms simultaneously permitted one waveform to be monitored while the other was undergoing modification.

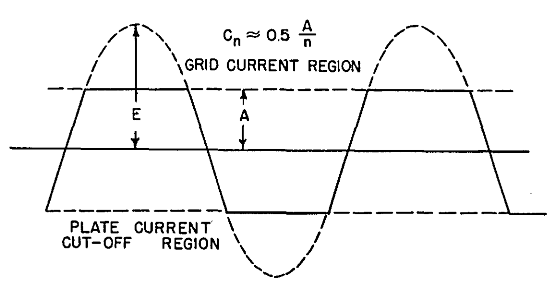

Fig. 2 - Definition of Symbols for Symmetrical Trapezoid Distortion

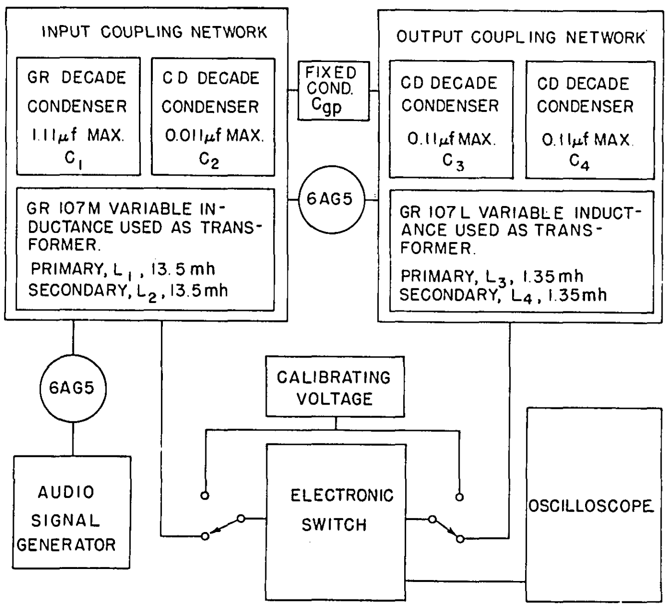

Fig. 3 - Block Diagram for Harmonic Generator Simulation Experiment

Photographs taken of some particular oscilloscope presentations are shown by Figures 5 - 11. The content of those photographs will be considered subsequently.

The two waveforms ordinarily observed were those from grid to ground, and from plate to ground of the pentode tube. It quickly became apparent that the loading provided by grid current flow was not large enough to produce the flat-topped distortion required. This situation can be improved either by raising the antiresonant impedance of the circuit L_2C_2, Figure 4, or by supplementing the non-linear loading effect arising from grid current flow. The latter condition can be realized by placing a 1N34 crystal from grid to ground, or from grid to cathode, with polarity arranged so that the crystal conducts when the grid conducts. By this means large distortion in the grid was secured. In fact, 1N34 crystals in parallel were tried with some increase in distortion as shown by some increase in output in the plate circuit. Since, however, the antiresonant impedance of L_2C_2, (Figure 4) is ordinarily 10 - 100 times the crystal impedance in the conducting direction, one crystal is ordinarily enough. The antiresonant impedance must be reduced as the resonant frequency is increased because inherently the L/C ratio must decrease. Consequently, either a lower-impedance crystal or crystals in parallel are required to get optimum distortion at high frequencies.

Initially it was felt that the grid circuit should be tuned to the driving frequency (that is, both L_1C_1 and L_2C_2, Figure 4, tuned to the input frequency) because this appeared to

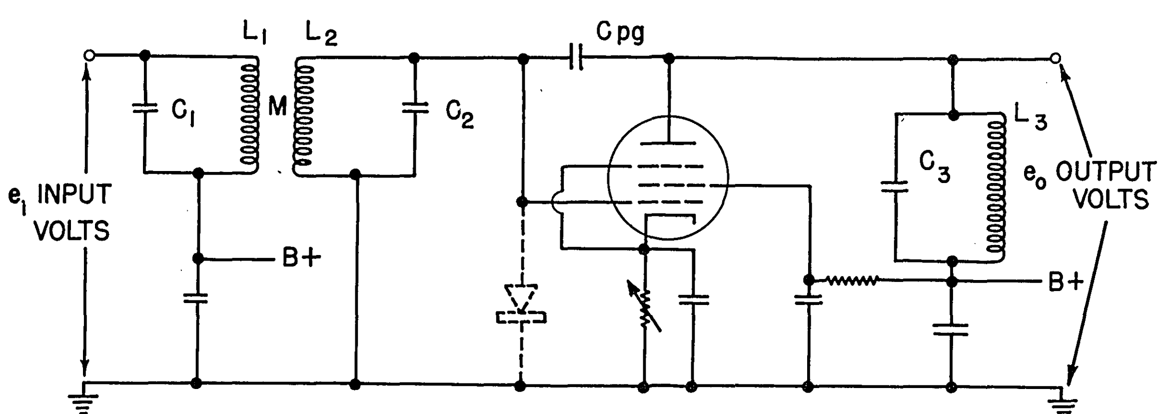

Fig. 4 - Circuit Used in Simulation Experiments
NOTE: See Table I for Important Component Sizes

be the obvious way to get a large driving voltage on the grid. The grid circuit was tuned to the incoming frequency while methods for producing sharp breaks in the grid waveform were experimentally investigated.

The use of 1N34 crystals in parallel with the control grid input impedance produced waveforms of the type shown in the oscillogram, Figure 5. The upper oscillogram shows the distorted grid waveform; the lower, the plate circuit waveform.

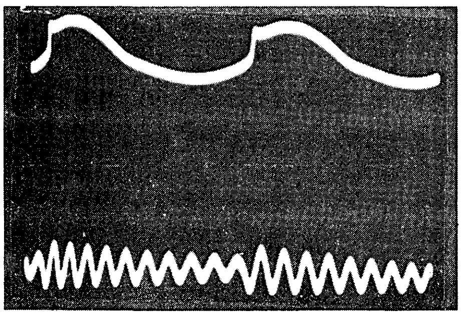

Fig. 5 - Oscillogram where L_2C_2 Is Tuned to the Input Frequency (See Fig. 2)

The distortion in the grid circuit shown by Figure 6 leaves something to be desired. The discontinuity establishes a transient of appreciable amplitude, but the low resonant frequency of the grid circuit leads to a rounded-off waveform not very rich in harmonics. From this transient standpoint the plate circuit oscillogram shows that there is only one discontinuity sharp enough to establish an appreciable transient.

If the circuit L_2C_2, Figure 4, is tuned to a higher frequency, transients richer in harmonics are established.

Figure 6 shows grid distortion obtained when the grid circuit is tuned to about five times the incoming frequency. The plate waveform has erratic changes in amplitude because the voltage swings in the grid circuit are too long in duration and phase relations frequently result in decreasing the amplitude in the plate circuit.

The oscillograms, Figures 7 through 9, show what happens when the grid circuit is tuned to the output frequency f_o, and the driving voltage is increased. Figures 7 and 8 show that the worthwhile increase in plate circuit amplitude comes in increasing the driving voltage e_i, Figure 4, from 1.0 to 2.0 volts rms. Further increase to 3.5 volts, Figure 9, yields diminishing returns. Above 3.5 volts input very little change in the plate circuit amplitude is secured.

In Figures 7 through 10 a transient is established which is superimposed upon a flat topped wave similar to the symmetrical trapezoid originally sought. As the plate waveforms show, throughout the ringing period of the grid circuit L_2C_2, Figure 4, the amplitude in the plate circuit builds up, whereas it dies down again during the flat top when the grid and crystal are conducting.

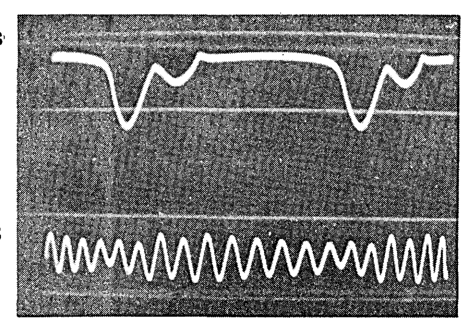

Fig. 6 - Grid Tuned to 5 Times Input Frequency

The increase in the amplitude due to ringing is 3 or 4 to 1 over the amplitude in Figures 5 and 6 with the same amplitude of drive measured across L_1C_1, Figure 4.

The frequency conversion gain obtained is affected to a degree by grid bias. However, if the distorting crystal is connected from grid to ground as shown in Figure 4, zero bias is always best for even harmonics. If the crystal is connected from grid to cathode, a fixed bias can be used thus reducing the plate current. A variable grid bias was used experimentally so that optimum adjustment could be made, but the output was not extremely critical to grid bias voltage.

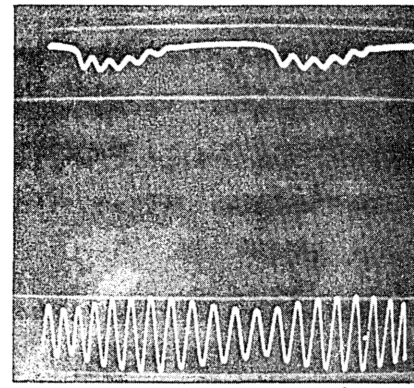

Fig. 7 - Grid Tuned to Output Frequency; e_i = 1.0 Volts RMS

Some frequency conversion gain is obtained with this multiplier stage as shown by Figures 7, 8, and 9. A 2.0-volt rms (2.8-volt peak) amplitude across L_1C_1 results in a voltage across L_3C_3 varying from a 5- to 8-volt peak. The faint horizontal line in these figures are voltage calibration lines. An input signal is thus converted to a signal at ten times the input frequency and about twice the input amplitude. The gain of the stage, G, measured by providing grid drive at the output frequency, was 30 in Figures 7, 8, and 9. It was estimated that a gain of 100 was required to get unity gain but this calculation did not include the contribution of the grid ringing to the Fourier Coefficient for the 10th harmonic. It is also to be noted that no load was coupled to L_3C_3. Adding the next distorter circuit coupled to L_3C_3 reduced the frequency conversion gain through the simulated stage to about unity. Increasing the Q of the L_3C_3 circuit can be easily accomplished in the unit required for multiplier range (b) to augment the gain if this is desired.

For the oscillogram, Figure 10, the distorting crystals were removed. Note that the flat top is almost lost, the ringing in the grid circuit is very much reduced, and the plate circuit amplitude is also much reduced.

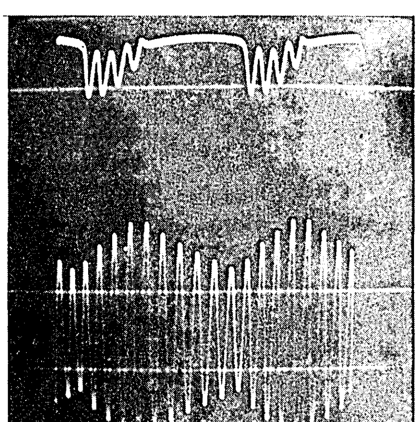

Fig. 8 - Grid Tuned to Output Frequency; e_i = 2.0 Volts RMS

LIMITATIONS IMPOSED BY INPUT CAPACITY OF THE STAGE

It is desirable to keep the ratio L_2/L_1 as large as possible so that the voltage applied to the grid may be large. At the same time L_1 should be made large so that the gain in the driving stage may be large. However, as was previously pointed out, L_2C_2 ought to be tuned to nf_i where n is the frequency multiplying factor and f_i the input frequency. In Table I the results of these conflicting requirements are summarized.

With an L_2/L_1 ratio of 1, the C_2/C_1 ratio is .01 for a frequency ratio $f_o/f_1 = 10$. Even in the 10 Mc output case, C_2 becomes very small. In one of the simulation experiments, C_2 was removed entirely with

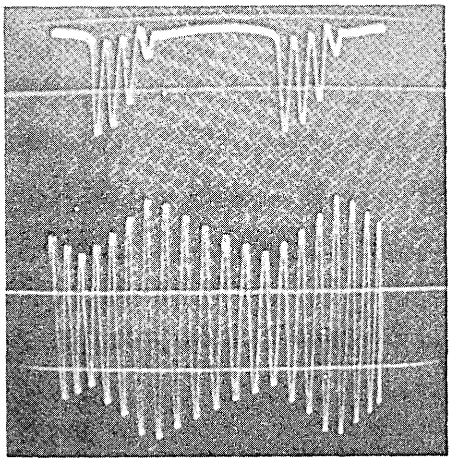

Fig. 9 - Grid Tuned to Output Frequency; e_i = 3.5 Volts RMS

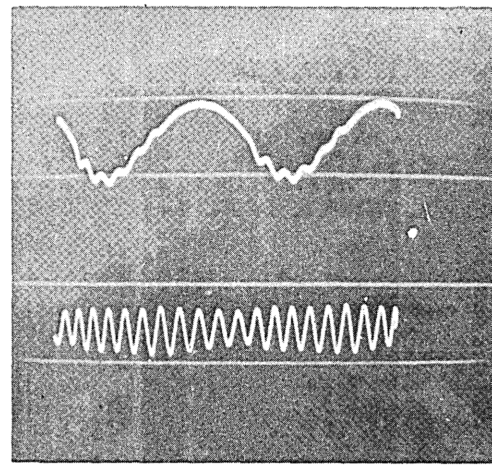

Fig. 10 - Grid Tuned to Output Frequency, 1N34 Crystal Removed; e_i = 2.0 Volts RMS

the result shown in Figure 11. Here the ringing of the coil L_2 at a higher frequency, its self-resonant frequency, is shown not to be particularly helpful. The original photograph barely resolves the damped sinusoid resulting from this action; consequently the reproduction is somewhat blurred right after the discontinuity in the grid waveform.

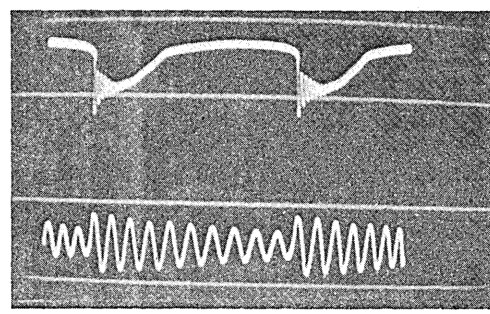

Fig. 11 - Capacitor C_2 Removed; e_i = 2.0 Volts RMS. Note Transient Due to Ringing of Coil at Its Self Resonant Frequency

A succeeding distorter circuit, say L_4C_4, coupled to L_3C_3 would require a tuning capacity of $C_4 = C_3/100 = 1.4\ \mu\mu f$. This is not obtainable with present miniature pentode tubes. The input capacity of the pentode tubes available is 3 or 4 times this figure. To resonant L_4C_4 to 100 Mc, L_4 must be reduced to about 1/3 or 1/4 of L_3. It is therefore necessary to accept either an unfavorable inductance ratio or a reduction in frequency conversion gain in the stage having the plate tank L_3C_3.

In cascading multiplier stages for which n = 10 serious limitations are encountered above about 30 Mc output.

FEEDBACK CONSIDERATIONS

A condenser, labelled Cpg in Figure 4, was placed in parallel with the plate to first grid capacity to simulate this important feedback component. Of course, the size to be chosen is not readily calculated since the actual Cpg existing in the final units is not known. The value taken, $.026\ \mu\mu f$ for 6AG5 tubes, is that given by one tube manual. The capacitor, Cpg, used for simulation was 760 (.026) = $20\ \mu\mu f$.

Regeneration was not large enough to be troublesome in simulating unit (b). In simulating unit (c) by the foregoing method, oscillation troubles were encountered. By decreasing the L3/C3 ratio, regeneration can be reduced (at a sacrifice in gain) so that self sustained oscillation does not occur. Pentode tubes rapidly lose their utility for this service above 10-Mc output, unless neutralization precautions are taken or unless locked oscillator types of multipliers are to be used. Locked oscillators are very practical for fixed-frequency applications particularly if a system devised by McCool at Naval Research Laboratory is used. In this system the plate voltage for the locked oscillator is obtained by rectifying the input signal. Consequently, if the driving signal is removed there is no output, but if the driving frequency shifts outside the locking range, output at an undesired frequency will be obtained.

The distortion method described in this paper is also useful for increasing the locking range of locked oscillators. The simulation apparatus can be used to study the effect of optimum distortion on locking range. A great many parameters are involved in specifying properly an operating condition, but a brief experiment along these lines was conducted. The presence of a crystal distorter expanded the locking range about 5 times in one particular experimental arrangement used.

COMMENTS ON MODULATION

In Problem A262R-U, it is important that modulation at the input frequency, f_i, present on the output, f_o, be small. In this respect large regeneration applied to the subject multiplier might be useful. Locked oscillators have inherently much less low-frequency modulation than that shown in Figures 5 through 11. On the other hand, simple filtering accomplishes acceptable performance.

CONCLUSIONS AND REMARKS

1. Simulation methods can be effectively used to study distorted waveforms in frequency multiplier circuits not ordinarily open to direct measurements means.

2. The conditions to be satisfied, so that a simulation circuit may reproduce the distorted waveforms in the grid circuit are, as defined by equations 1-7 under Theory of Simulation:

(a) $\dot{I}_{1s} = I_{os}$ (complex steady state currents equal)

(b) $E_{ot} = E_{1t}$ (complex voltage equal at beginning of transient)

(c) $G_1 = G_o$ (conductance of distorting component duplicated)

(d) $$\frac{C_o}{C_1} = \frac{L_o}{L_1} = \frac{t_o}{t_1} = \frac{\omega_1}{\omega_o} = \frac{f_1}{f_o}$$

3. Experimental results on frequency multipliers can be obtained much more rapidly by using simulation methods than the corresponding transient problem can be solved analytically.

[4] U.S. Patent 2377894 - "Automatic control for locked oscillators" by William A. McCool, 12 June 1945.

4. A non-linear impedance needs to be used in parallel with the grid so that large transients may be established in the grid circuit of the multiplier. Rectifier crystals, such as the 1N34, are excellent for this purpose.

5. For optimum performance the grid- and plate-tuned circuits L_2C_2 and L_3C_3 should be tuned to the same frequency. Above 30 Mc, excessive regeneration may dictate that either the harmonic generator grid and plate circuits be tuned to different frequencies or neutralizing methods applied.

6. A grid drive of 2 or 3 volts rms is required for sharp cut-off pentode tubes of the 6AG5 type. Below a 1.0-volt drive, the frequency conversion gain decreases rapidly. Above a 3-volt drive the output changes very little.

7. Frequency multipliers of the subject type can be built having an output to input frequency ratio of 10 with an accompanying amplitude ratio of 1 below about 30 megacycles. (The power gains, in circuits where the impedance across which the voltages are measured is about 20,000 ohms, can be made unity below 30 Mc.)

8. Tube capacities limit the frequency range over which unity gain in frequency-conversion operation can be obtained. It is estimated that locked-oscillator techniques would have to be used above 30 Mc if unity gain is to be realized for a frequency multiplication of 10.

9. The distortion technique here described may be utilized to increase the locking range of locked oscillator systems used as frequency multipliers.

RECOMMENDATIONS

1. The distortion techniques herein described should be applied to lower voltage-gain stages such as grounded grid amplifiers. Information from this development is required to complete the design of unit (d) and possible unit (c).

2. An attempt should be made to combine the locked oscillator method used by McCool with the distortion method outlined here to obtain a multiplier which cannot go into self oscillation when the input frequency is varied.

* * *

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