A Double-Stage Injection-Locked Oscillator for Optically Fed Phased Array Antennas

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Abstract —**In an optically fed phased array antenna system, the microwave carrier signal is transmitted via a modulated lightwave to each active T/R (transmit/receive) module, where it must be converted back to the microwave domain. Currently, efficient optical to microwave conversion is extremely difficult, as the detected microwave signal is weak and noisy. A novel circuit, containing a high-gain/low-noise microwave injection-locked oscillator, has been developed to improve the interface between the optical and microwave components. The circuit utilizes two FET's and a dielectric resonator, which serves as a frequency-dependent feedback element. The circuit, designed to operate at about 8 GHz, provides significant amplitude and phase noise suppression. In addition, the circuit realization is compatible with MMIC technology.**

I. Introduction

A new generation of phase array antennas will utilize a large number of individually powered monolithic microwave integrated circuit (MMIC) transmit/receive (T/R) modules. The frequency and phase synchronization of each T/R module is critical in ensuring array coherence. In an optically fed phased array antenna system, coherence is accomplished by the distribution of the frequency reference signal to each T/R module via high-speed fiber-optic links [1]–[3]. The reference signal is utilized for both up- and down-conversion of information at the T/R module [3] and must have adequate phase stability, good noise performance, and sufficient power to drive the mixers of the T/R module. The modulated optical signal, which carries the RF reference, is weak and noisy, particularly at higher frequencies [4]. Therefore, special circuitry is required to provide an efficient interface between the optical feed and the T/R module microwave circuits. The primary purpose of this circuitry is to convert the optical signal into a strong, low-noise microwave signal.

A block diagram of the reference circuit, including an injection-locked oscillator for an optically fed phased array antenna, is shown in Fig. 1. The frequency reference signal is provided from the CPU (central processing unit) to the T/R level through a fiber-optic network. The high-speed intensity

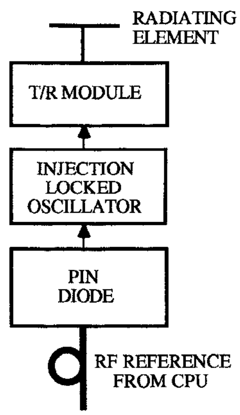

Fig. 1. Block diagram of RF reference circuit including the injection-locked oscillator.

modulation of the light may be performed by either a direct modulation of the laser or an external electro-optic modulator. The microwave reference signal is then directly detected at each element by a highly sensitive p-i-n photodiode. Next, the power level of the recovered microwave reference signal has to be enhanced. For this purpose, a novel circuit has been designed and built utilizing an injection-locked oscillator to serve as an interface between the optical feed and the T/R module microwave circuitry. The injection-locked mode of operation has been chosen because its AM noise suppression is higher than that of the amplifier mode of operation [5], [6]. The circuit design criteria are high gain, high frequency and phase stability, high AM (amplitude modulation) and FM (frequency modulation) noise compression, and compatibility with MMIC implementation.

II. Circuit Description

A schematic block diagram of the new injection-locked oscillator is shown in Fig. 2. The circuit utilizes two transistors, which are operated as wide-band amplifiers. The frequency reference used for injection locking is input to the oscillator from port 1. Port 2 is the output of the injection-locked oscillator, which is fed to the local oscillator (LO) port of the T/R level mixers. The dielectric disk resonator serves as a narrow-band feedback element which determines the free-running oscillation frequency. The function of the matching circuits is to provide efficient coupling between the p-i-n photodetector and the input of the oscillator, as well as

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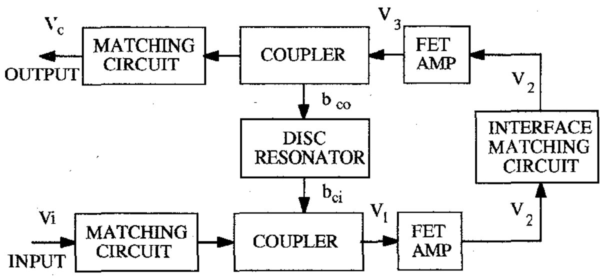

Fig. 2. Two-stage injection-locked oscillator circuit.

between the oscillator output and the T/R module mixers. It should be noted that efficient matching of the p-i-n to the first transistor requires special consideration [7].

Several advantages are realized in the present design configuration. The two-stage approach enhances the output power of the oscillator while providing good noise performance even at low injection levels. To minimize the noise of the oscillator, a low-noise FET or HEMT is used in the first stage. This stage is operated as an almost linear amplifier, with its bias adjusted according to low noise requirements. In the second stage, a high-gain FET is operated in a nonlinear mode to increase the oscillator output power level and also to compress the amplitude of oscillation. In addition, for a specific locking band, the two-stage design, because of its increased gain, requires an injected signal power lower than that of a single-stage design.

The new arrangement also offers advantages in the oscillator noise performance. The present design provides significant AM compression because of the enhanced gain of the two-stage injection-locked oscillator. In addition, both amplifiers are designed to operate over a much wider band than the locking band; therefore, they will not exhibit frequency dependence over the locking range. The oscillation frequency is therefore controlled merely by the resonant frequency of the dielectric resonator.

High frequency and phase stability can also be expected over a wide temperature range owing to the inherent temperature stability of the dielectric resonator [8]. The phase shift of an injection-locked oscillator is a function of the difference between the injected frequency and the free-running oscillation frequency [9]. Because of manufacturing tolerances and differential temperature effects, that difference will vary from element to element in a phased array system. The manufacturing tolerances can be compensated by appropriate adjustments of the specific oscillators. Furthermore, the use of a dielectric resonator with a low temperature coefficient minimizes the temperature dependence of the free-running oscillation frequency and hence the temperature dependence of the phase shift. This feature is very important because temperature-dependent phase shift errors can cause severe beam steering errors in phased array antennas [2].

The present design requires only standard MMIC transistors and passive components (capacitors and no inductors); therefore, it is totally compatible with MMIC technology. In addition, since the injection-locked oscillator is not a reflection type, no circulator will be required. A nonlinear circuit analysis of the new injection-locked oscillator along with experimental results will be presented in the following sections.

III. Analysis of the Double-Stage Injection-Locked Oscillator

The injection-locked mode of operation has been investigated by numerous authors. Several papers cover the principles of injection locking [10]–[15]. Other papers discuss the most important properties of injection-locked oscillators [16]–[18]. However, there are some requirements which can not be met applying the well-known single-stage circuit constructions. For example, in many cases the locking band and the Q factor are determined by the specification of the circuit; therefore the gain cannot be chosen freely. This difficulty may be overcome in part by the double-stage construction suggested in this paper.

In this analysis the capabilities of the double-stage injection-locked oscillator construction have been evaluated. A nonlinear circuit analysis has been performed to determine the expected circuit performance. Both the free-running and injection-locked modes of operation has been investigated, specifically the AM compression and AM-to-PM conversion. The optimum operating point for the circuit with respect to noise performance [5] has been determined.

A. Basic Relationships

The simplified block diagram shown in Fig. 2 is used for the analysis. Voltage control is assumed in the analysis consistent with the FET implementation such that only the fundamental voltage appears at each port [9]. The port voltages of the amplifier stages are designated as V_1, V_2, and V_3 and all are assumed to be real quantities. Furthermore, b_{ci} is the voltage coupling factor at the input, b_{co} is the voltage coupling factor at the output, and Q_e is the external Q factor of the dielectric disk resonator, where the loss of the dielectric resonator is neglected in the feedback circuit.

Both amplifier stages are considered to be nonlinear for completeness and are described by the following voltage transfer functions, where only two terms of the general power series presentation of nonlinear networks are retained for simplicity:

$$V_2 = A_{10}(1 - n_{13}V_1^2)V_1 \tag{1}$$

$$V_3 = A_{20}(1 - n_{23}V_2^2)V_2. \tag{2}$$

A_{10} and A_{20} are the voltage gains for the first and second stages, respectively. In addition, n_{13} and n_{23} are the coefficients of the third power terms of the power series for the first and second stages, respectively.

The overall voltage transfer function for the two amplifiers may now be obtained from the previous expressions:

$$V_3 = A_0(1 - n_3V_1^2)V_1 \tag{3}$$

where

$$A_0 = A_{10}A_{20} \tag{4}$$

$$n_3 = A_{20}n_{13} + A_{10}^3 n_{23}. \tag{5}$$

The voltage transfer function of the dielectric resonator is also required and is given as follows:

$$F = F_r + jF_i = \frac{1 - 2jQ_e\delta}{1 + (2Q_e\delta)^2} \tag{6}$$

where F_r is the real part and F_i is the imaginary part of the transfer function; Q_e is the external Q factor of the resonator; and δ is the relative frequency deviation from the resonant frequency f_0 as defined below:

$$\delta = \frac{f - f_0}{f_0}. \tag{7}$$

In general, the voltage at port 1 consists of both feedback and injected voltage components and may be expressed as

$$V_1 = V_3 b_c (F_r + jF_i) + V_{ir} + jV_{ii} \tag{8}$$

where b_c is the overall voltage feedback coupling factor, which is simply the product of the input and output coupling factors:

$$b_c = b_{ci} b_{co}. \tag{9}$$

V_{ir} and V_{ii} are the real and imaginary parts of the injected signal voltage, and V_3 is given in (3). Equation (8) is valid for operation in the locking band, where the operating frequency is determined by the frequency of the injected signal.

B. Free-Running Oscillation

First, the case of free-running oscillation is considered. The free-running case is characterized by the absence of the injected signal, such that $V_{ir} = V_{ii} = 0$ and the free-running oscillation frequency is equal to the resonant frequency of the dielectric resonator. Substituting V_1 from (8) into (3), the voltage V_{30} is expressed as

$$V_{30}^2 = \frac{A_0 b_c - 1}{A_0 n_3 b_c^3} \tag{10}$$

from which the condition of oscillation is obtained:

$$b_c > \frac{1}{A_0}. \tag{11}$$

The feedback coupling factor must be greater than the reciprocal of the voltage gain for oscillation.

The free-running oscillator power may now be found and is given by

$$P_{30} = \frac{1}{2} G_L V_{30}^2 = \frac{1}{2} G_L \frac{A_0 b_c - 1}{A_0 n_3 b_c^3}. \tag{12}$$

An expression for maximum power can be obtained by differentiating P_{30} with respect to b_c and equating the derivative to zero. An optimum feedback coupling factor, $b_{c\,\mathrm{opt}}$, is then obtained:

$$b_{c\,\mathrm{opt}} = \frac{3}{2A_0} \tag{13}$$

and the maximum output power can be expressed by

$$P_{30\max} = \frac{2}{27} G_L \frac{A_0^2}{n_3} \tag{14}$$

where the voltage belonging to the maximum output power is given by

$$V_{30m}^2 = \frac{4}{27} \frac{A_0^2}{n_3}. \tag{15}$$

The maximum output power of the free-running oscillation will serve as a reference. Both input and output powers in the injection-locked mode of operation will be normalized to that value.

C. Injection Locking

Next, the injection-locked mode of operation at the fundamental frequency is considered where the injected voltage is a complex quantity as previously mentioned. The real part of (8) may be substituted into (3) to obtain a third-order expression for V_3:

$$V_3^3(n_3 A_0 b_c^3 F_r^3) + V_3^2(3n_3 A_0 b_c^2 F_r^2 V_{ir}) + V_3(3n_3 A_0 b_c F_r V_{ir}^2 - b_c F_r A_0 + 1) - A_0 V_{ir} + n_3 A_0 V_{ir}^3 = 0. \tag{16}$$

The phase difference, θ, between the injected signal and the feedback signal may now be found by solving the imaginary part of (8):

$$V_{ii} = -b_c V_3 F_i \tag{17}$$

so that

$$\Theta = \tan^{-1}\left(\frac{-b_c V_3 F_i}{V_{ir}}\right). \tag{18}$$

Finally the powers are normalized to the maximum output power of the free-running oscillator. The normalized input and output powers are denoted by small letters: p_i and p_3 or p_o. The output power, p_o, is lower than p_3; however, this difference will be omitted.

1) Locking Band: The locking band can be determined based on (8). If the phase shift, θ, is equal to $\pm\pi/2$, then the real part of the injected voltage, V_{ir}, will be zero and the voltage V_3 will be the same as in the case of the free-running oscillator (eq. (10)). The imaginary part of (8) utilizing (6) then becomes

$$V_3 b_c \frac{2Q_e\delta}{1 + (2Q_e\delta)^2} - V_{ii} = 0. \tag{19}$$

Solving the previous expression yields the following expression for the locking band:

$$B_L \approx \frac{f_0}{Q_e b_c}\sqrt{\frac{p_i}{p_o}} \tag{20}$$

The locking band is inversely proportional to the feedback coupling factor, b_c, which is again inversely proportional to the voltage gain A_0. An increased locking band may therefore be obtained by increasing the voltage gain and properly choosing b_c. The advantage of the increased gain capability of the double-stage injection-locked oscillator is thus obvious with respect to the locking band.

2) Noise Properties: In the analysis, the injected signal is considered as having mainly amplitude (AM) noise. The AM noise can be described by the amplitude modulation of the input signal; this approximation is valid for a narrow frequency band around the carrier [19]. Thus the AM noise can be considered a perturbation of the signal, and it is affected mainly by the so-called derivative nonlinear transfer characteristics: the AM compression and the AM-to-PM conversion [9]. These properties have been investigated in detail.

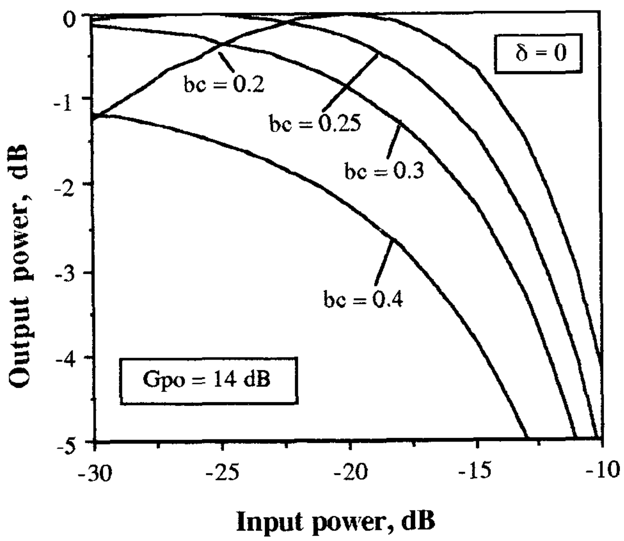

Fig. 3. Oscillator's normalized output power versus normalized input power.

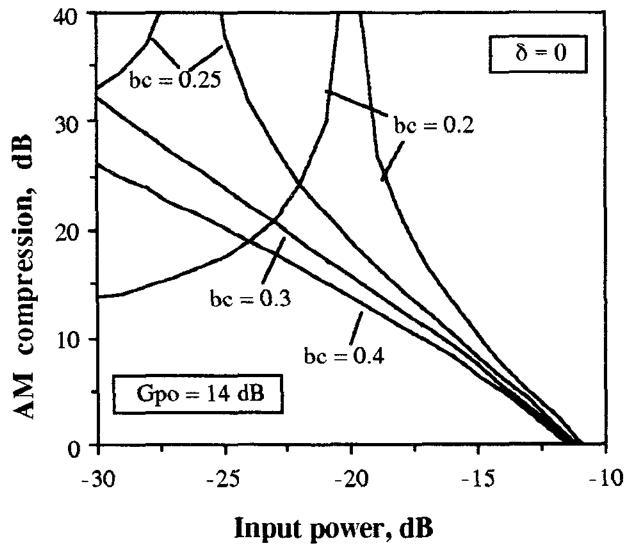

Fig. 4. AM compression versus normalized input power.

Owing to the nonlinearity of the circuit, the AM of the transmitted signal is changed—usually reduced—which is called AM compression. Another property of the nonlinear circuit is the AM-to-PM conversion, which means that phase modulation (PM) of the transmitted signal will appear to the output when an amplitude-modulated signal is applied at the input. In other words, the AM at the input is some extent converted to PM at the output.

Specifically, the AM compression (cp) and the AM-to-PM conversion (cv) of the double-stage injection-locked oscillator are given by the following expressions, respectively:

$$\mathrm{cp} = 20\log\left|\frac{p_o}{p_i}\frac{dp_i}{dp_o}\right| \ [\mathrm{dB}] \tag{21}$$

$$\mathrm{cv} = 0.26\frac{180}{\pi}p_i\frac{d\theta}{dp_i} \ [^\circ/\mathrm{dB}]. \tag{22}$$

The AM compression of the circuit reduces the AM noise. However, there is a noise conversion effect as well. Because of the AM-to-PM conversion, a part of the AM noise is converted to phase noise as well. For an optimum solution, therefore, a high AM compression and an associated low AM-to-PM conversion are needed; these are functions of the voltage gain, feedback coupling factor, and injected power. These relationships are numerically analyzed and the results are discussed next.

IV. Results of the Analysis

First, the output power was investigated as a function of the input power for different feedback coupling factors. This is plotted in Fig. 3. Here the small-signal power gain, 14 dB, corresponds to the designed circuit gain. Thus, maximum output power is obtained with $b_c = 0.3$ in the free-running mode of operation. In the injection-locked mode of operation, the maximum output power is the same as in the free-running mode of operation. However, in the injection-locked mode of operation, a different feedback coupling factor belongs to the maximum output power. For a specific input power, an optimum feedback coupling factor can be obtained which gives maximum output power. At this point, the AM compression has a pole (or infinite value). For example, for a normalized input power of -20 dB, the maximum output power is obtained with an overall feedback coupling factor b_c equal to 0.2.

Next, the AM compression was investigated as a function of the input power for different feedback coupling factors. This is depicted in Fig. 4. Here the power gain is again 14 dB. The maximum output power is obtained with $b_c = 0.3$ in the free-running mode of operation. As Fig. 4 shows, a higher AM compression can be reached if the overall feedback coupling factor b_c is smaller than 0.3. For example, for a normalized input power of -20 dB, an infinite AM compression is obtained if $b_c = 0.2$. When the gain is high (i.e., the input power is low), the AM compression is very high. With decreasing input power, both the AM compression and the gain are increased. Thus, the two-stage injection-locked oscillator simultaneously exhibits a high AM compression and a high gain. Therefore, the AM noise originating in the modulated laser diode and in the fiber-optic link will be significantly suppressed, while the associated gain will be high.

Fig. 5 shows the AM compression as a function of the relative frequency deviation if the overall feedback coupling factor b_c is equal to 0.2. For an input power of -20 dB a very high AM compression is obtained in a wide band. The curve exhibits two poles.

Another property of the circuit is the AM-to-PM conversion, which also has an effect on the noise performance. Fig. 6 shows the AM-to-PM conversion as a function of relative frequency deviation for different input powers. In this case, the small-signal power gain, G_{p0} is again 14 dB. This figure reveals that the magnitude of the AM-to-PM conversion increases with a decrease in the input power. When the AM noise is compressed significantly, therefore, the magnitude of the AM-to-PM conversion increases, and a part of the AM noise is converted into phase noise. However, the AM-to-PM conversion of the two-stage injection-locked oscillator is low enough; thus, there is no noticeable noise conversion. Furthermore, as seen in Fig. 6, an optimum adjustment can be found where the AM-to-PM conversion is zero. Consequently, at this operation point, the AM noise will not be converted to phase noise. This is a very convenient feature of injection-locked oscillators.

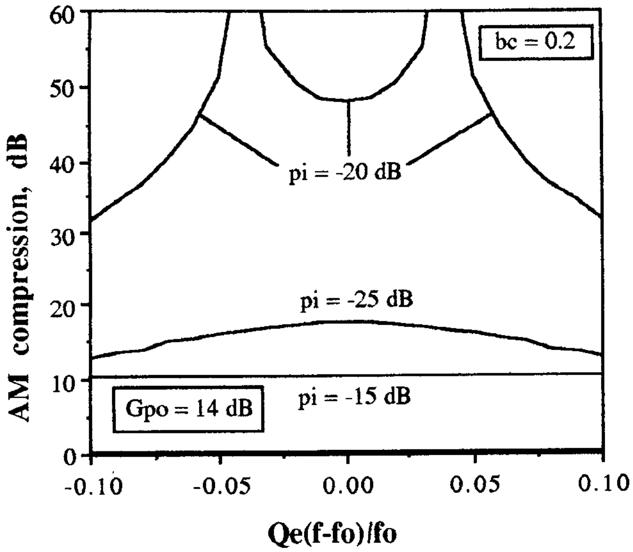

Fig. 5. AM compression versus relative frequency deviation for different normalized input powers.

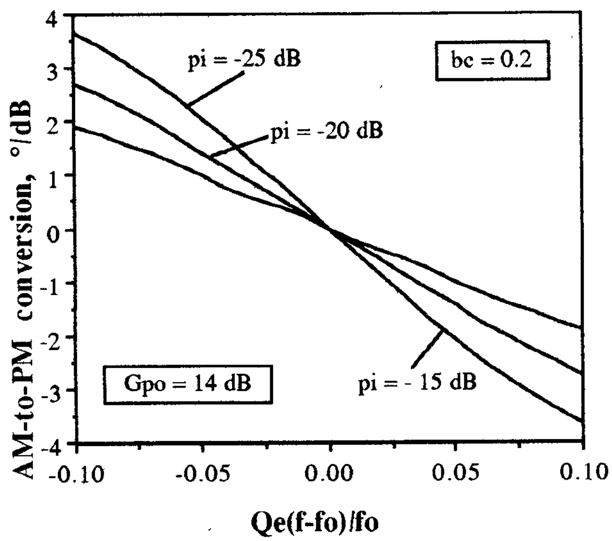

Fig. 6. AM-to-PM conversion response for different normalized input powers as a function of frequency deviation.

For the reference signal, a low phase noise is also required. The phase noise contribution of the fiber-optic link is not negligible; therefore, it should be suppressed as well. The phase noise can also be reduced significantly by the two-stage injection-locked oscillator. For this purpose, a resonator with a high Q factor is applied at the input of the injection-locked oscillator, very effectively reducing the phase noise of the reference signal by its high selectivity. However, the temperature dependence of the oscillator is enhanced if its Q factor is increased. This contradictory problem has been solved by applying a dielectric resonator which exhibits a high Q factor and a very low temperature coefficient simultaneously [20].

V. Experimental Results

The two-stage injection-locked oscillator (ILO) has been built in a hybrid MIC construction to operate around 8 GHz. Because matching to the p-i-n photodiode was not considered in this effort, the circuit was designed with standard 50 Ω terminations and connections. A circuit simulation was performed using microwave CAD Touchstone and more than 14 dB of power gain was predicted for the two stages. The circuit layout is shown (although not to scale) in Fig. 7. Two general-purpose packaged FET's were utilized for both stage 1 and stage 2. Although these devices will not provide as low a noise figure for stage 1 as an HEMT, they would satisfy the requirements. For the feedback circuit, a V-shaped microstrip line is applied which makes possible the variation of the coupling factor. The position of the dielectric resonator is movable, allowing the coupling factor and the distance from the stages to be adjusted.

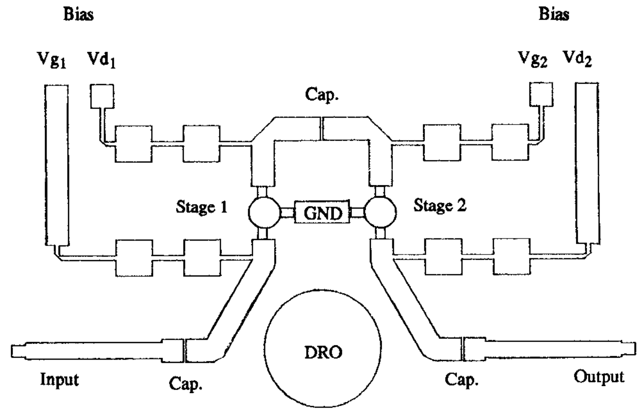

Fig. 7. Circuit layout.

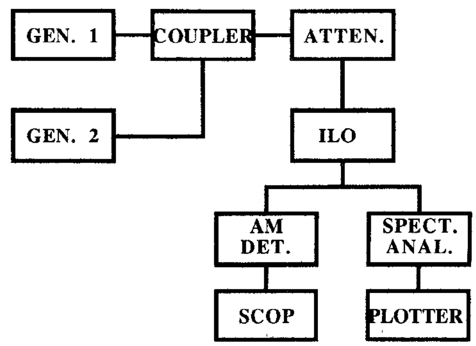

Fig. 8. Measurement setup.

Experimental investigations have been carried out utilizing the developed circuit. The oscillation frequency was around 8.5 GHz, and the output power was approximately 1 mW if the cable losses are deducted. The position of the dielectric disk resonator was optimized, resulting in an overall coupling factor of 0.2 approximately. The Q factor of the dielectric resonator was 4000.

The experimental setup is shown in Fig. 8. At the input two signals have been combined by a directional coupler. The level difference between them was set to 20 dB, and the input level of the circuit under test has been changed by a variable attenuator, keeping the level difference of the two signals at the same value. The frequency separation between these signal was always smaller than half of the locking band. The frequency difference has been varied from 20 kHz to 200 kHz. A coaxial amplitude detector and a spectrum analyzer have been used to evaluate the transfer property of

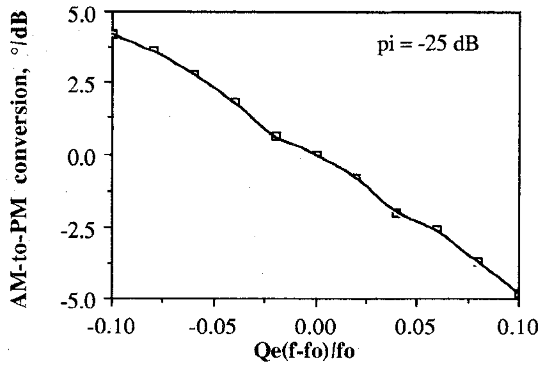

Fig. 9. Measured AM-to-PM conversion versus relative frequency deviation.

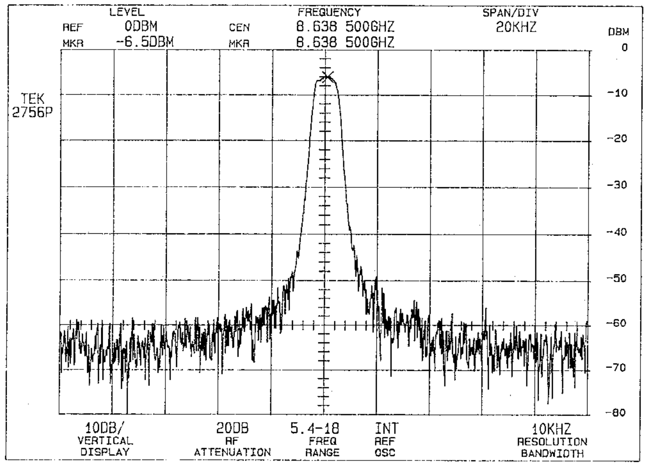

Fig. 10. Noise spectrum.

the ILO. The combined input signal means simultaneous amplitude and phase modulation. In this way the amplitude and phase noise have been simulated [9].

In the first measurement, the AM compression was investigated. First the power level of the signal injected at the input was varied in a 10 dB range between −25 dB and −15 dB relative to the 1 mW output power. No significant change has been observed in the output power over this range, indicating that the AM compression is very high. Further, the coaxial amplitude detector was used for measuring the AM component of the combined signal at both input and output. The detected signal at the input of the ILO served as a reference. The level of the input signal was varied in a 10 dB range between −25 dB and −15 dB. At the output of the ILO, the detected signal level was always more than 15 dB below the level of the reference signal. Thus the AM compression was always higher than 15 dB, which is in a good agreement with the analytical results.

If the AM compression is high enough, the AM modulation is practically eliminated, and the spectrum of the output signal should contain two sidebands with equal amplitudes. In this case, the level of the sidebands at the output has to be 3 dB less than that at the input. If the suppression of the sidebands is higher than 3 dB, the phase modulation has also been reduced. The sideband levels have been checked by the spectrum analyzer, and they were equal and 28.4 dB below the carrier level. This means that the phase modulation has been decreased by 5.4 dB because of the filtering effect of the dielectric resonator at the input of the circuit. This effect was not dependent on the input signal level; thus, it was the result of the linear transfer property of the circuit. Consequently, beside a more than 15 dB reduction in the amplitude noise, a further 5.4 dB decrease in the phase noise can be obtained by the application of the new two-stage injection-locked oscillator.

In the next experiment, the AM-to-PM conversion was measured. The result is plotted in Fig. 9, which shows the AM-to-PM conversion as a function of relative frequency deviation for an input power of −25 dB. Finally, the noise spectrum has been measured and plotted as shown in Fig. 10. In this case, the resolution bandwidth of the spectrum analyzer was 10 kHz. As can be seen, the signal is very clean: its noise bandwidth is around 10 kHz. The average level of the noise is approximately −60 dBc at 40 kHz or more from the carrier.

VI. Conclusion

The novel circuit, containing a high-gain, low-noise microwave injection-locked oscillator, clearly improves the interface between the optical and microwave subassemblies of optically fed phased array antennas. The circuit, which utilizes two FET's and a dielectric resonator, provides significant amplitude and phase noise suppression and has been designed to be compatible with MMIC technology.

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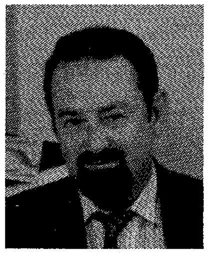
Tibor Berceli (SM'77) was born in Budapest, Hungary. He received the Dipl. Ing. and Dr. Ing. degrees, both in electrical engineering, from the Technical University of Budapest, Hungary, in 1951 and 1961, respectively. From the Hungarian Academy of Sciences, he received the Candidate of Technical Science degree on the basis of his dissertation on surface wave low-loss transmission lines in 1955, and the Doctor of Technical Science degree on the basis of a dissertation on the linearization of reflex klystron modulators in 1965.

He Berceli joined the TKI, Research Institute for Telecommunications, Budapest, in 1951. Since that time he has been involved in microwave research and development work. He has investigated surface wave transmission lines, dielectric waveguides, traveling wave amplifiers, reflex klystron oscillators, several kinds of microwave semiconductor oscillators and amplifiers, parametric circuits, up- and down-converters, injection-locked oscillators, etc. His present field of interest is the optical–microwave interaction. Since 1962 Dr. Berceli has also been associated with the Technical University of Budapest as a Professor of Electrical Engineering. He has taught courses on microwave techniques, active nonlinear microwave circuits, and ratio communications systems.

Dr. Berceli is the author of 82 papers and six books published in English. He has presented 56 papers at international conferences. He was a visiting professor at the Polytechnic Institute of Brooklyn, Brooklyn, NY, in 1965, at University College London in 1986, and at Drexel University, Philadelphia, PA, for the 1988–89 academic year. Dr. Berceli served as Chairman of the Organizing Committee for the URSI International Symposium on Electromagnetic Theory in 1986. He is currently the Conference Chairman of the 20th European Microwave Conference.

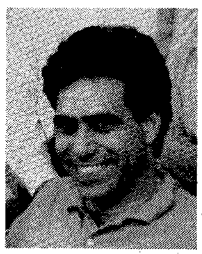
William D. Jemison (S'83–M'85) received the B.S.E.E. degree from Lafayette College, Easton, PA, in 1985 and the MESc degree from Pennsylvania State University in 1988. He is currently with the Department of Electrical and Computer Engineering at Drexel University, Philadelphia, PA, working toward the Ph.D. degree. His research interests lie primarily in optically controlled antennas and microwave devices.



Peter R. Herczfeld (S'66–M'67–SM'89–F'91) was born in Budapest, Hungary, in 1936. A U.S. citizen, he received the B.S. degree in physics from Colorado State University in 1961. He then received the M.S. degree in physics in 1963 and the Ph.D. degree in electrical engineering in 1967 from the University of Minnesota. Since 1967 he has been on the faculty of Drexel University, Philadelphia, PA, where he is a Professor of Electrical and Computer Engineering.

He has published over 180 papers in solid-state electronics, microwaves, photonics, solar energy, and biomedical engineering. He has served as project director for over 50 projects. Dr. Herczfeld has taught 20 different courses at the graduate and the undergraduate level and has lectured extensively in the U.S. and in ten other countries. He is currently Director of the Center for Microwave–Lightwave Engineering at Drexel.

Dr. Herczfeld is a member of SPIE and the ISEC. A recipient of several research and publication awards, including the Microwave Prize, he also served as guest editor for a special issue of the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES and the JOURNAL OF LIGHTWAVE TECHNOLOGY entitled Applications of Lightwave Technology to Microwave Devices, Circuits, and Systems.



Afshin S. Daryoush (S'84–M'86) was born in Iran in 1957. He received the B.S. degree in electrical engineering in 1981 from Case Western Reserve University, Cleveland, OH. He then received the M.S. degree in 1984 and the Ph.D. degree in 1986 from Drexel University, Philadelphia, PA, also in electrical engineering.

After graduation he joined the staff of Drexel University, first as Research Assistant Professor and then, beginning in 1987, as DuPont Assistant Professor of Electrical and Computer Engineering. In 1989 he was promoted to Associate Professor. He has conducted research in the area of optically controlled microwave devices and subsystems, high-speed fiber-optic links, and system studies of large-aperture phased array antennas. During the summers of 1987 and 1988, he was a Summer Faculty Fellow at NASA, Lewis Research Center, Cleveland, OH, conducting research on high-speed fiber-optic links for ACTS project. In the summers of 1989 and 1990 as a Summer Faculty Fellow, he conducted research on high-speed LED for 1.25 Gb/s fiber-optic links for computer backplanes at the Naval Air Development Center, Warminster, PA.

Dr. Daryoush has authored or coauthored over 90 technical publications in the areas of light interaction with passive and active microwave devices, circuits, and systems. He has lectured frequently at workshops and international symposia. The recipient of the Microwave Prize from the 16th European Microwave Conference, Dublin, Ireland, he also received the best paper award at the IMPATT Session of the 1986 International Microwave Symposium, Baltimore, MD. He has also been awarded a U.S. patent on the optically controlled patch antenna. He is a member of Sigma Xi.

Arthur Paolella (M'89) was born in Atlantic City, NJ, in 1957. He received the B.S. degree from Monmouth College in electronic engineering in 1982 and the M.S. degree from Fairleigh Dickinson University in electrical engineering in 1985. He is currently a Ph.D. candidate at Drexel University, Philadelphia, PA, where he is performing research at the Center for Microwave/Lightwave Engineering.

At present, he leads the Microwave Photonics Team within the Microwave/Lightwave Branch of the Electronics Technology and Devices Laboratory at Ft. Monmouth, NJ. He is currently conducting research in the area of optical control of GaAs MMIC's, and MESFET optical detectors and is working on the development of analog microwave fiber-optic links. In 1987 he was selected by the Army's Professional Long Term Training Program to attend Drexel University. He has authored or coauthored several publications on optical control of GaAs MMIC's and on advanced millimeter-wave Gunn oscillators. He is the holder of two patents.